

RADIO TRANSMITTERS

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PREFACE

There is a wealth of published information covering the circuitry used in radio transmitters. Similarly there are a number of books devoted entirely to the subject of modulation, oscillators, measurements, etc. However, there has been a lack of books which present the transmitter engineer or operator with a comprehensive discussion of those areas which are of immediate interest to him, and which are arranged to present this material from the "transmitter" viewpoint. It is the purpose of this book to collate this material which is of particular interest to transmitter design engineers and which will be useful to those engaged in transmitter operation and maintenance. Certain portions should also be of interest to the advanced amateur.

It is assumed that the reader has at least attained the necessary knowledge required to obtain a commercial operator's license. Therefore no attempt has been made to go into the elementary principles of electricity. Conversely, the book does not give mathematical derivations of the equations presented throughout the text. The general scope of the work does not permit this. Rather the book is intended to present a practical analysis of transmitter operation in all its various phases, together with usable information pertaining to specific problems in transmitter operation. Thus for those whose problems demand a more complete analysis or additional mathematical support, the reference lists provided at the end of each chapter will indicate additional source material.

The analysis of transmitter operation has been developed by first grouping and discussing circuits and components common to all transmitting equipment. Chapter 12 is devoted to the discussion of the particular characteristics of transmitters in specific services. Measurement techniques particularly applicable to transmitters are described in Chap. 13, and finally a brief summary of hazards associated with transmitters is given in Chap. 14.

We wish to thank the many manufacturers and organizations who contributed material. We also wish to thank ITT Laboratories for permission to include some previously unpublished material. Acknowledgment of the sources of illustrations has been given individually.

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CONTENTS

<i>Preface</i>	v
Chapter 1. Introduction	1
1-1. Definition	1
1-2. Licensing and Identification	1
1-3. Classification of Transmitters	2
1-4. Signal-transmission Terms	2
1-5. Transmission Frequencies	3
1-6. Radiated Power Levels	4
1-7. Modulation	5
1-8. Power Supplies, Control Circuits, and Cooling	7
Chapter 2. Frequency-control Techniques	9
2-1. Basic Methods of Control	9
2-2. Inductance-Capacitance Oscillators	10
2-3. Quartz Crystals	12
2-4. Crystal Mountings	14
2-5. Crystal Ovens	16
2-6. Crystal-frequency Correlation	16
2-7. Military Standard Crystal Units	16
2-8. Crystal-oscillator Circuits	19
2-9. Parallel-resonant Circuits	20
2-10. Series-resonant Circuits	20
2-11. Drive Level	21
2-12. Variable-frequency Crystal Oscillators	22
2-13. Frequency Stability of Crystal Oscillators	23
2-14. Frequency Control in F-M Transmitters	24
2-15. Frequency Control of Microwave Oscillators	25
2-16. Frequency Synthesizers	27
2-17. Frequency Multipliers and Dividers	27
2-18. Mixers	28
2-19. Filters	29
2-20. Drift-canceled Oscillator	30
2-21. Synthesizing Systems	32
Chapter 3. R-F Power Amplifiers	37
3-1. General Amplifier Considerations	37
3-2. Amplifier Classifications	37
3-3. Dynamic Operating Conditions	38
3-4. Amplifier Configurations	38

3-5. Amplifier Efficiency	39
3-6. Efficiency and Plate-current Conduction Angle	39
3-7. The Characteristics of the Class A Amplifier	40
3-8. The Characteristics of the Class B Amplifier	43
3-9. The Characteristics of the Class C Amplifier	46
3-10. A Simplified Method of Amplifier Design	47
3-11. The Grounded-cathode Amplifier	54
3-12. The Grounded-grid Amplifier, General Characteristics	54
3-13. The Grounded-plate Amplifier	59
3-14. Amplifier Output Load Resistance	60
3-15. Amplifier Stabilization	61
3-16. The Linear Amplifier	72
3-17. Wideband Amplifier	80
3-18. Frequency Multipliers	82
3-19. Practical Circuit Considerations	84
Chapter 4. Power Tubes	89
4-1. Transmitting-tube Construction	89
4-2. The Cathode	89
4-3. Grids	93
4-4. Anodes	93
4-5. General Types of Transmitting Tubes	95
4-6. Klystron Amplifier Tubes	95
4-7. Reflex Klystrons	99
4-8. Traveling-wave Tubes	100
4-9. Magnetrons	103
4-10. Platinotrons, Amplitrons, and Stabilotrons	104
Chapter 5. Coupling Circuits	108
5-1. Coupling Networks	108
5-2. Parallel-tuned Circuits	108
5-3. Series-tuned Circuits	111
5-4. The Double-tuned Circuit	112
5-5. Envelope Delay	115
5-6. Matching Network Configurations	115
5-7. The Basis of the L, π , and T Matching Networks	116
5-8. The L Network	117
5-9. The π Network	118
5-10. The T Network	120
5-11. The π -L Network	121
5-12. Reactive Load Impedances	122
5-13. Link-coupled Circuits	122
5-14. Balanced-to-unbalanced Coupling Networks	125
5-15. Matching-network Efficiency	128
5-16. Impedance Matching in Transmission Lines	129
5-17. Impedance Matching in Waveguide Circuits	132
Chapter 6. Amplitude Modulation	134
6-1. Amplitude Modulation	134
6-2. Sidebands	134

6-3. Double-sideband Amplitude Modulation	136
6-4. On-off Keying	137
6-5. Diode Modulator	139
6-6. Plate Modulation	141
6-7. Grid Modulation	146
6-8. Amplification of Modulated R-F Voltages	152
6-9. Carrier-suppression Techniques	152
6-10. Vestigial-sideband Transmission	153
6-11. Phase-to-amplitude Modulation	154
6-12. Negative Feedback	155
6-13. Single Sideband	157
6-14. Advantages of Single Sideband	157
6-15. Methods of Generating an SSB Signal	158
6-16. The Balanced Modulator	162
6-17. Quartz Crystal Filters	166
6-18. Mechanical Filters	169
6-19. Audio Phase-shift Networks	170
6-20. R-F Phase-shift Networks	171
6-21. Frequency Translation in SSB Transmitters	172
6-22. Linearity Measurement in SSB Transmitters	174
6-23. Signal-to-distortion Ratio	177
6-24. Sideband Suppression	178
 Chapter 7. Angle and Pulse Modulation	 181
7-1. Angle Modulation	181
7-2. Sidebands Produced by Angular Modulation	182
7-3. Sideband Characteristics	183
7-4. P-M to F-M Conversion	184
7-5. Methods of Generating Phase Modulation	186
7-6. Direct Generation of Frequency Modulation	189
7-7. Pulse Modulation	193
7-8. Sidebands Produced by Pulse Modulation	194
7-9. Pulse Generators	195
7-10. Pulse-amplitude Modulation	198
7-11. Pulse-duration Modulation	199
7-12. Pulse-position Modulation	200
7-13. Pulse-code Modulation	200
7-14. Delta Modulation	201
 Chapter 8. Power Supplies	 206
8-1. A-C Voltage Regulators	206
8-2. Power-supply Voltage Control	213
8-3. Power-supply Circuits	215
8-4. Rectifiers	218
8-5. Power-supply Filter Circuits	224
8-6. Transformers and Chokes	228
8-7. Corona	231
8-8. Filament Starting	232
8-9. D-C Regulators	236
8-10. D-C to D-C Power Supplies	238

Chapter 9. Control and Protective Circuits.	244
9-1. Control and Protective Circuits	244
9-2. Functions of the Control Circuits	244
9-3. Power Control	244
9-4. Transmitter Sequencing and Interlocking Circuits	247
9-5. Indication and Alarms	249
9-6. Transmitter Recycling Circuits	251
9-7. Remote Control	254
9-8. Transmitter Telemetry	257
9-9. Automatic Tuning of Transmitters	259
9-10. Protective-circuit Functions	260
9-11. Protective-circuit Philosophy	260
9-12. The Fuse	262
9-13. The Circuit Breaker	263
9-14. Circuit-breaker Applications	263
9-15. Electronic Crowbar	265
9-16. R-F Overload Protection	266
9-17. Across-the-line Diagrams	268
Chapter 10. Cooling	270
10-1. Transmitter Cooling	270
10-2. Forced-air Cooling of Tubes	270
10-3. Liquid Cooling of Tubes	277
10-4. Tube Temperature Measurement	281
10-5. General Cabinet Cooling	281
Chapter 11. R-F Components	283
11-1. R-F Components	283
11-2. Directional Couplers	283
11-3. Dummy Loads	290
11-4. Lossy Dielectric Loads	291
11-5. R-F Filters	292
11-6. Vestigial-sideband Filter	293
11-7. Diplexers	294
11-8. The Magic T	297
11-9. The Coaxial Hybrid Ring	298
11-10. Common Frequency Diplexing of Transmitters	298
11-11. Special Adjustment of Combined Transmitters	300
11-12. Power Dividers	301
11-13. Transmission-line Switching	303
11-14. Microwave Ferrites	305
11-15. Isolators	305
11-16. Circulators	306
Chapter 12. Transmitter Characteristics	309
12-1. Transmitter Applications	309
12-2. Navigation	309
12-3. Radar	310
12-4. Telegraph Transmitters	312
12-5. Telephone Transmitters	316
12-6. Mobile Transmitters	317

CONTENTS

xi

12-7. Telemetry Transmitters	318
12-8. Broadcast Transmitters	319
12-9. Television Transmitters	320
12-10. Broadcast Repeaters and Translators	328
12-11. R-F Generation for Noncommunication Purposes	329
Chapter 13. Transmitter Measurement Techniques	333
13-1. Transmitter Measurements	333
13-2. Power Measurements	333
13-3. Frequency Measurements	344
13-4. Time Measurements	359
13-5. A-M and F-M Measurements.	361
13-6. Demodulated Audio Measurements	367
13-7. Telegraph Keying Measurements.	372
13-8. Broadband Radio System Measurements	373
13-9. Pulse-transmitter Modulation Measurements	375
13-10. Monochrome Video Modulation Measurements.	375
13-11. Color Video Measurements	388
13-12. Antenna Measurements	393
13-13. Spurious Transmitter Output Measurements	405
Chapter 14. Hazards Associated with Transmitters	417
14-1. Health Hazards.	417
14-2. Electric Shock	417
14-3. Treatment of Electric Shock	418
14-4. Precautions against Electric Shock	419
14-5. Lightning Hazards	421
14-6. Power-line Shorts	421
14-7. X-ray and R-F Radiation Hazards	422
14-8. Chemical Hazards	426
<i>Appendix</i>	429
<i>Index</i>	451

INTRODUCTION

1-1. Definition. Radio transmitters, for the purpose of discussion in this book are defined as generators of useful radio-frequency (r-f) energy which are intended to be connected to some form of electromagnetic radiator. The "receiver" of the transmitted energy may be only a few inches away as in the case of induction heating equipment or a diathermy machine, or it may be thousands of miles away.

A transmitter is an assembly of an r-f generator, amplifiers, modulator, power supplies, control circuits, and cooling equipment. Each of these items will be discussed in detail.

1-2. Licensing and Identification. The operation of a transmitter generally requires a license issued by the government having jurisdiction over the location or vehicle in which the transmitter is located. In the United States the Federal Communications Commission (FCC) specifies a value of radiation above which a transmitter license is required. Below this value the equipment is called an *incidental or restricted-radiation device*. For frequencies above 1,600 kc the radiation field must not exceed 15 μv per m at a distance of

$$\frac{157}{f_{\text{kc}}} \text{ ft}$$

The limiting values for frequencies between 10 and 1,600 kc are given in Table 1-1. No radiation is permitted on the international distress frequency which is 500 kc.

TABLE 1-1. FIELD STRENGTH LIMITS FOR UNLICENSED TRANSMITTERS

Frequency, kc	Distance, ft	Maximum field strength, $\mu\text{v}/\text{m}$
10—490	1,000	$\frac{2,400}{f_{\text{kc}}}$
510—1,600	100	$\frac{24,000}{f_{\text{kc}}}$

These limits permit the use of "wireless" microphones and radio-control systems without a license.

Radio transmissions are identified by call letters which are assigned when a license is issued. Blocks of call letters are assigned to the various countries by international agreement. A list of call-letter assignments is given in Table A-1 (in the Appendix). Transmitter operation in the United States is governed by the various rules and regulations issued by the FCC (1).

1-3. Classification of Transmitters. For licensing purposes transmitters are classified according to the bandwidth, the type of modulation, and the type of intelligence being sent. The classification is identified by combinations of numerals and letters as shown in Table 1-2.

TABLE 1-2. TRANSMITTER EMISSION CLASSIFICATIONS

Example: 3 A 3 a = ssb telephony			
Bandwidth, kc	Type of modulation or emission	Type of information	Supplementary characteristics
System bandwidth required for proper reproduction at the receiver. Examples A3a voice—3 kc A3 voice—6 kc A5 VSB television—6,000 kc For frequencies below 10 kc, a maximum of two significant figures may be given after decimal: Examples A1 telegraphy, 25 wpm—0.1 kc F1 telegraphy, 100 wpm—1.25 kc	A—Continuous wave including all types of a-m B—Damped wave (spark) F—Angular modulation (frequency or phase) P—Pulsed emission	0—No keying or modulation 1—Telegraphy without other modulation 2—Telegraphy using modulated carrier. (Carrier may or may not be keyed) 3—Telephony 4—Facsimile 5—Television 6—Multiplex telegraphy 9—Composite and cases not covered above	No letter given for full carrier a-m or any type of f-m a—Single-sideband, reduced carrier (ssb) b—Independent sideband, reduced carrier (isb) c—Other types of reduced carrier d—Pulse amplitude modulation (PAM) e—Pulse width modulation (PWM) f—Pulse position modulation (PPM)

1-4. Signal-transmission Terms. The bandwidth required for a system will depend upon the rate at which information is to be sent, the type of modulation, and the amount of distortion that can be tolerated

in the reproduced signal. The International Radio Consultative Committee (CCIR) which is a body of the International Telecommunication Union (ITU) makes recommendations concerning the necessary bandwidth for various types of transmission (2).

In the case of telegraph transmission the rate of sending information is given in *bauds*. In on-off telegraphy, the on, key-down, or radiated-power condition is called the *mark*, while the off condition is called the *space* condition. In frequency-shift transmission it is conventional to call the *higher* frequency the mark. The speed in bauds is the total number of elementary marks and spaces per second. The speed in bauds will be *twice* the number of cycles per second (cps).

In the case of the transmission of business-machine or telemetered data, it is more usual to express the speed in *bits* or *kilobits* (1,000 bits) per second. Bits per second and bauds represent equivalent speeds.

The information rates for telephony, facsimile, and television are simply given in cycles per second.

Table 1-3 gives conversion factors to convert words per minute into bauds or cycles per second for various codes (3).

TABLE 1-3. FACTORS TO CONVERT WPM TO BAUDS OR CPS FOR VARIOUS CODES

Code	bauds	cps
International Morse.....	0.9	0.45
American Morse.....	0.85	0.425
Cable Morse.....	0.4	0.2
Start-stop teleprinter.....	0.742	0.371
Synchronous teleprinter (no start or stop pulses).....	0.50	0.25

1-5. Transmission Frequencies. The frequencies which are allocated to transmitters in the various classes of service are constantly changed as the state of the art advances, and no attempt will be made in this book to give the specific frequency allocations. The general division of the frequency spectrum is decided at the international telecommunication conventions which are held approximately at 10-year intervals. Frequency allocations in the United States are given in vol. 2, part 2 of the FCC "Rules and Regulations" (1).

The frequencies of fixed stations throughout the world are recorded by the International Frequency Registration Board (IFRB), which is another body of the ITU, and the information is published in the "Radio Frequency Record."

The useful r-f spectrum is divided into ranges which are given names and band numbers. The band number refers to the power of 10 appli-

cable to the high end of each range. Designations given by the CCIR are given in Table 1-4.

TABLE 1-4. CCIR FREQUENCY-RANGE DESIGNATIONS

Band number	Frequency range	Designation
4	3-30 kc	vlf, very low
5	30-300 kc	l-f, low
6	300-3,000 kc	m-f, medium
7	3,000-30,000 kc	h-f, high
8	30-300 Mc	vhf, very high
9	300-3,000 Mc	uhf, ultrahigh
10	3-30 kMc	shf, superhigh
11	30-300 kMc	ehf, extremely high
12	300-3,000 kMc	Not named

The uhf and shf bands are further broken down into lettered designations which are given in Table 12-1.

The allowable tolerance or variation from assigned frequencies is also established by international agreement. In some cases the individual governments may enforce more restrictive tolerances. Observance of the tolerance restrictions is an important factor in permitting the maximum number of transmitters to occupy the frequency spectrum without mutual interference.

Methods of frequency control are discussed in detail in Chap. 2.

1-6. Radiated Power Levels. The maximum amount of radiated power permitted for each class of service is specified in government regulations. In many cases the particular license for a transmitting station may further limit the radiated power. In some cases, such as in the amateur service and maritime service where it is difficult to standardize antenna efficiencies, transmitter power is limited on the basis of input to the final amplifier rather than radiated power. In general it is the policy to limit power levels to the minimum required to perform the desired service.

Methods of r-f power amplification are discussed in Chap. 3. Chapter 4 covers power-amplifier tubes, and Chap. 5 discusses methods of coupling power amplifiers to antennas. In Chap. 11 a number of special components which are associated with the problem of coupling power to antennas at very high frequencies and microwave frequencies are described.

It is important that 99 per cent or more of the radiated power lies within the bandwidth permitted for each class of service. *Spurious* out-of-band radiation can seriously interfere with other services. The

most common type of spurious radiation is on frequencies which are multiples or *harmonics* of the desired frequency.

Another type of spurious radiation is the excessive sidebands generated as the result of overmodulation. In a transmitter where the output frequency is derived from a mixer, the unwanted mixer products must be sufficiently attenuated. Similarly, in cases where the output frequency is derived from multipliers, the subharmonics must not be radiated.

When there is some degree of coupling between two antennas connected to a pair of transmitters operating at one location, *intermodulation* can occur in each transmitter. For example, if one transmitter is on frequency A and the other is on a higher frequency B , the first transmitter can generate spurious frequencies at $A - (B - A)$, $A \pm 2(B - A)$, $A \pm 3(B - A)$, etc. Similarly, the B transmitter can generate spurious outputs at $B + (B - A)$, $B \pm 2(B - A)$, etc.

Spurious output frequencies can be attenuated by filters, stubs, or other devices in the transmission line between the transmitter and antenna. It is important that direct radiation of spurious frequencies from the cabinet is kept to a minimum.

1-7. Modulation. Intelligence is imparted to a radio transmitter by the process of *modulation*. This is a process whereby the characteristics of a wave are modified by means of another, lower-frequency wave. A basic example is the human voice. The mouth cannot be moved at a rate of more than about 10 cps, while the aperture of the human ear is too small to allow reception of this frequency. The vocal chords can generate higher frequencies which the ear can hear, but these frequencies are only converted into intelligence when modulated by the mouth and tongue. The higher frequency is called the *carrier* wave.

Two or more channels of information can be sent on a single carrier by modulating the carrier with *subcarriers* which, in turn, are modulated with the information to be sent. The type of modulation applied to the carrier may be different from that applied to the subcarriers. The standard notation which indicates modulation types shows the modulations in turn from left to right as the information proceeds to the carrier. For example, a-m/f-m¹ means that the information to be sent amplitude-modulates the subcarriers, and the subcarriers frequency-modulate the carrier.

The principal types of modulation are *amplitude*, *angular*, and *pulse*. Amplitude-modulation methods are taken up in Chap. 6. Angular- (frequency- and phase-) and pulse-modulation methods are described in Chap. 7.

No modulation method is optimum for all conditions, and many factors must be considered (4). The choice of the best method for a given appli-

¹ Sometimes written as AM/FM.

cation will depend on a number of system considerations which are beyond the scope of this book, but the relative merits of a few types will be mentioned. Modulation should not only be considered as a means of superimposing a message onto a carrier but should also be considered as a means of protecting the message from mutilation by noise during transmission.

The performance of a communication system is usually expressed in terms of the *signal-to-noise ratio* (s/n or snr) at the equipment output

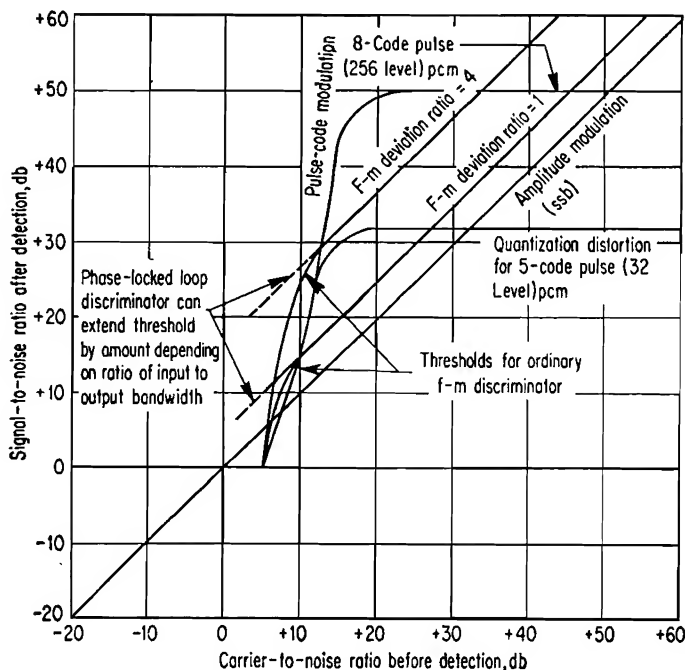


FIG. 1-1. Comparison of modulation types.

terminals. There will be a *carrier-to-noise* ratio at the input to the receiver detector, which will depend on the power being received from the transmitter and the noise power in the bandwidth which is passed by the receiver filters. In the case of amplitude modulation the demodulated signal-to-noise ratio will be no better than the carrier-to-noise ratio. In the case of frequency modulation, on the other hand, there will be an *improvement*, provided the signal exceeds a certain *threshold*, that is proportional to the modulation index.¹ The threshold level can be reduced by means of *phase-lock* or *synchronous* detection techniques at the receiver (5).

¹ See Sec. 7-1 for definition of modulation index.

A more rapid improvement in the output signal-to-noise ratio can be achieved by the use of certain types of pulse modulation. Figure 1-1 shows typical values of signal-to-noise ratio after detection as related to carrier-to-noise ratios before detection for frequency modulation and also for pulse-code modulation. The noise in each case is that measured over the bandwidth being used. For random noise, the noise power will be proportional to the bandwidth in which the measurement is taken.

Other methods of improving receiver output signal-to-noise ratios for a given transmitter power include the use of preemphasis (accentuation of high frequencies) in the transmitter with subsequent deemphasis in the receiver. Also peak compression at the transmitter with subsequent expansion in the receiver (compandor) can be used. These techniques will not be discussed as they are not related to transmitter design other than the selection of the output power for a given application. The methods of establishing necessary transmitter power outputs for a required type of service involve many systems considerations and will not be discussed (6).

1-8. Power Supplies, Control Circuits, and Cooling. A transmitter could be described as a piece of hardware which converts a-c or d-c power into r-f power. The a-c or d-c input voltage must usually be converted to higher and lower voltages for proper operation of the various circuits in a transmitter, and Chap. 8 describes methods of doing this. Control and protection of power supplies and other circuits is discussed in Chap. 9.

A certain amount of power is always lost in the form of heat in power supplies and amplifiers. Cooling methods are discussed in Chap. 10.

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1. Federal Communications Commission: "Rules and Regulations," Superintendent of Documents, Washington D.C. Vol. I: part 0, Statement of Organization, Delegations of Authority and Other Information; part 1, Practice and Procedure; part 13, Commercial Radio Operators; part 17, Construction, Marking and Lighting of Antenna Structures. Vol. II: part 2, Frequency Allocations and Radio Treaty Matters; General Rules and Regulations; part 5, Experimental Radio Services (other than broadcast); part 15, Incidental and Restricted Radiation Devices; part 18, Industrial, Scientific and Medical Service. Vol. III: part 3, Radio Broadcast Services; part 4, Experimental, Auxiliary and Special Broadcast Services. Vol. IV: part 7, Stations on Land in Maritime Services; part 8, Stations on Shipboard in the Maritime Services; part 14, Public Fixed Stations and Stations of the Maritime Services in Alaska. Vol. V: part 9, Aviation Services; part 10, Public Safety Radio Services; part 11, Industrial Radio Services; part 16, Land Transportation Radio Services; Vol. VI: part 12, Amateur Radio Service; part 19, Citizens Radio Services; part 20, Disaster Communications Service. Vol. VII: part 6, International Fixed Public Radio-communication Services; part 21, Domestic Public Radio Services (other than maritime mobile); (*Note: Volumes VIII, IX, and X relate to accounts, reports, and tariffs and do not cover equipment.*)

2. "Reference Data for Radio Engineers," 4th ed., pp. 21-23, International Telephone & Telegraph Corporation, 1956.
3. See Ref. 2, p. 846.
4. Kelley, G. J.: Choosing the Optimum Type of Modulation: A comparison of several communication systems, *IRE Trans. on Commun. Systems*, vol. CS-6, pp. 14-21, June, 1958.
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FREQUENCY-CONTROL TECHNIQUES

2-1. Basic Methods of Control. The importance of the stability of the output frequency of a radio transmitter was discussed in Chap. 1. The following methods of transmitter frequency control are used:

1. Tuned circuit using lumped inductance and capacitance, transmission line, or cavity
2. Quartz crystal resonator
3. Frequency synthesizer using crystals or tuning fork as basic stable source

Simple tuned-circuit control is useful where equipment is under constant direct control of an operator and where adequate monitoring facilities are available to check that the transmitter output frequency is what is desired. This type of control is in wide use by amateur operators where flexibility is desirable and only short-term stability is needed. High-frequency broadcast transmitters are also frequently controlled by this method. With careful design and control of operating conditions, relatively good stabilities are obtainable. Short-term (1-hr) stabilities of 1 part in 10^6 and medium-term (1 day) stabilities of 10 parts in 10^6 have been reported (1).

Crystal Control. Excellent frequency stability is obtainable by directly controlling an oscillator with a quartz crystal. This is the typical method of control for fixed-frequency operation such as broadcasting.

Automatic Frequency Control. In many cases, direct frequency control is less practical than a servomechanism type of control where a constantly developed correction voltage is derived from a frequency or phase comparator. This correction voltage is then applied to a circuit which adjusts the oscillator frequency. One example is in microwave equipment where direct-frequency multiplication from a stable source is not economical at driving-power levels. Another example is a reactance-tube-modulated f-m transmitter.

Synthesizers. Where a large number of stable output frequencies are required from a transmitter, it becomes impractical to supply a separate crystal for each frequency. A unit called a *frequency synthesizer* will

generate a large number of stable frequencies from one or a few stable sources.

With the increased use of single-sideband operation, close control of both receiver and transmitter are necessary. High-stability synthesizers make this type of operation practical in mobile units (2,3). Another reason for the increased need for synthesizers is that calling and answering procedures are time-consuming and can be avoided by setting the transmitter and receiver to the correct frequency before any transmission takes place.

2-2. Inductance-Capacitance Oscillators. The stability of simple inductance-capacitance oscillators depends on the extent with which the tuned circuit can be isolated from the associated tube or transistor. It also depends on how well the tuned-circuit components can be made to maintain their values independently of environmental conditions. Isolation of the tuned circuit from the rest of the equipment is achieved by shielding. Isolation of the tuned circuit from the driving tube can be accomplished by using the

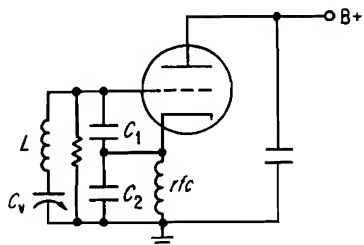


FIG. 2-1. Clapp oscillator.

low-impedance series-tuned oscillator circuit developed by Gouriet and Clapp (4,5). The basic circuit is shown in Fig. 2-1. The minimum condition for oscillation is given by

$$g_m = \frac{2\pi f C_1 C_2}{Q C_v} \quad (2-1)$$

where g_m = transconductance, μhos

f = operating frequency, Mc

C_1 = grid-to-cathode capacity, $\mu\mu\text{f}$

C_2 = cathode-to-ground capacity, $\mu\mu\text{f}$

Q = Q of tuned circuit which is usually Q of the coil alone in this circuit

C_v = capacity in tuned circuit, $\mu\mu\text{f}$

The circuit will oscillate if g_m is equal to or greater than the above quantity. The best stability against tube voltage variations is obtained when C_1 and C_2 have maximum values consistent with continuing oscillation. It can be seen from Eq. (2-1) that C_1 and C_2 may be increased if Q or g_m is increased. The g_m required to maintain oscillation increases as the operating frequency is increased, and this type of circuit is not generally used over a frequency range of more than 1.2 to 1. This ratio is adequate for high-frequency broadcasting, and amateur use where continuous coverage is only needed over restricted ranges.

For highest stability of LC oscillators, low-level operation with some form of automatic level control is desirable. This can be done by rectifying a portion of the output and using the direct current obtained to control the oscillator grid-bias voltage.

Temperature Compensation. The tuned-circuit components should be stabilized against temperature variations. One method is to install the tuned circuit in an oven and control the temperature. With the circuit shown in Fig. 2-1, the tuned circuit may be separated from the associated tube or other heat-producing components and connected with a coaxial cable. Capacities C_1 and C_2 will usually be much larger than the capacity of a convenient length of coaxial cable. The tuned-circuit components will then remain close to room temperature, and, if the room-temperature changes are small, compensation will not be necessary.

If operation over a wide temperature range is necessary, compensation of the inductance, capacitance, or both may be needed (6). If the changes are small, the frequency change is approximately given by

$$-\frac{\Delta f}{f} = \frac{\Delta C}{2C} + \frac{\Delta L}{2L} \quad (2-2)$$

where Δf , ΔC , and ΔL are the changes in frequency, capacity, and inductance over the temperature range being considered. f , C , and L must be expressed in the same units. $\Delta C/C$ and $\Delta L/L$ are called the temperature coefficients and are usually expressed in *parts per million per degree centigrade* (ppm per $^{\circ}\text{C}$). Fixed ceramic capacitors are available in a variety of temperature coefficients from +150 ppm per $^{\circ}\text{C}$ to -1,600 ppm per $^{\circ}\text{C}$. Temperature-compensating inductances are not generally used but can be made by using a variable iron core which is adjusted by some thermally actuated device. A single-turn loop, coupled to the inductor and rotated in response to temperature changes has also been used. In a variable-frequency oscillator where both the capacitor and inductor have temperature coefficients, exact compensation is only obtainable at one frequency. Where the requirements do not justify the use of a temperature-controlling oven, the best compromise is to select a low-coefficient fixed inductance and then to select the temperature coefficients of C_1 and C_2 , considering the circuit of Fig. 2-1, so that a zero frequency-temperature coefficient is obtained at mid-frequency in the tuning range. It is usual practice to make the temperature-compensating capacitor a small percentage of the total capacity and then to use a relatively high temperature coefficient such as 750 ppm per $^{\circ}\text{C}$. The main section of the capacitor should be a stable type such as silvered mica.

Other Circuits. There are many varieties of LC circuits which are covered in detail in a number of textbooks (7). Simplified circuits of the

Hartley (tapped inductance), Colpitts (tapped capacitor), electron-coupled, and tuned-plate tuned-grid oscillators are shown in Fig. 2-2. In the Hartley and Colpitts oscillators the effect of the tube on the tuned circuit can be reduced by the insertion of small reactances between the tuned circuit and the tube (8). In the vhf range the tuned circuits may be transmission lines, while in the uhf range the tuned circuit may be a

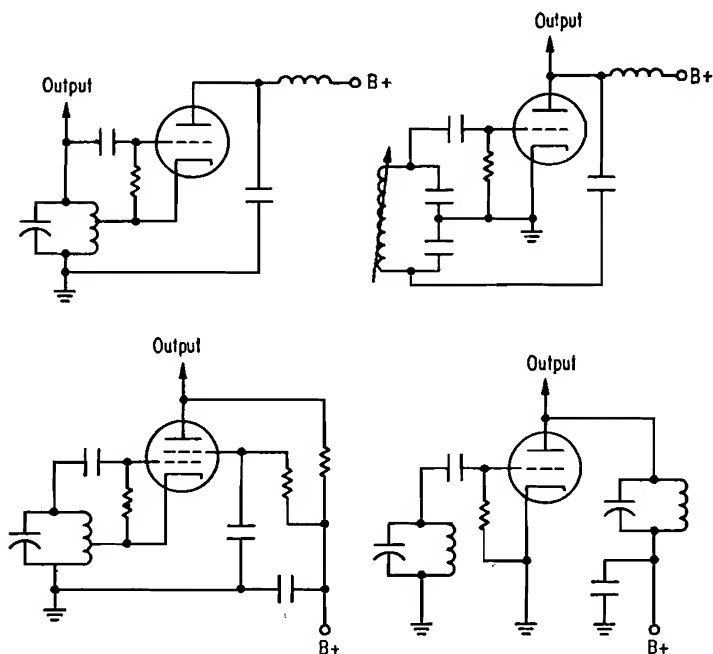


FIG. 2-2. (Top left) Hartley oscillator. (Top right) Colpitts oscillator. (Bottom left) Electron-coupled oscillator. (Bottom right) Tuned-grid-tuned-plate oscillator.

resonant cavity (9,10). A *traveling-wave tube* which is a broadband high-gain amplifying tube may be used for a relatively stable uhf oscillator by using a high- Q resonant cavity in the feedback path (11).

2-3. Quartz Crystals. Quartz, which is silicon dioxide, is the commonest of minerals, but in its ordinary form it shows no *piezoelectric* properties. Silicon atoms have positive charges, and oxygen atoms have negative charges. In an ordinary piece of quartz in which the molecules are randomly arranged, physical pressure will move the molecules to new positions in a random manner, and no net change in electric charge between opposite sides occurs. In crystalline quartz, however, the molecules and atoms are arranged in exact symmetry, and, if physical pressure is applied which causes deformation along the *mechanical axis*, an electric charge will be observed between faces which are along the *electrical axis*.

Conversely, if a voltage is applied to these faces, a physical deformation along the mechanical axis will occur. This is known as the *piezoelectric effect*, and many detailed discussions of the phenomenon will be found in the literature (12,13). Quartz plates will vibrate in synchronism with the frequency of an applied voltage but will vibrate at a vastly increased amplitude when the applied frequency corresponds with the natural mechanical resonant frequency of the plate. Properly designed crystal-

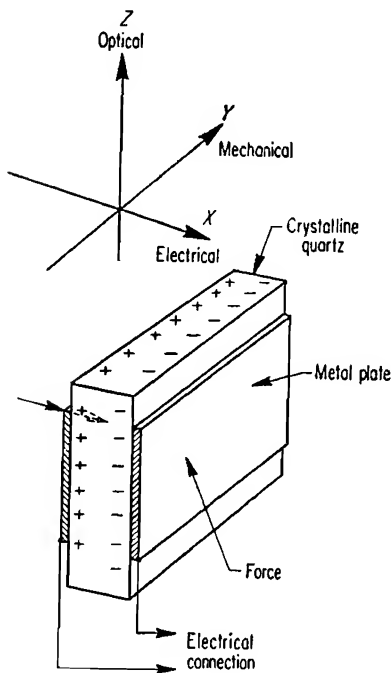


FIG. 2-3. Transforming energy. (*The Scientific Monthly*.)

controlled oscillators will oscillate only at the mechanically resonant frequency or its multiples which are called *overtones*. Figure 2-3 shows the three axes and the basic operation of a quartz crystal.

Crystal Cuts. The physical properties of a quartz crystal vary in different directions through its structure. The natural resonant frequency

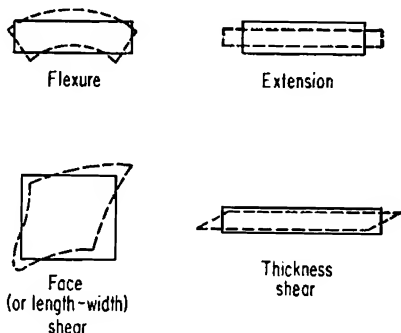


FIG. 2-4. Useful fundamental modes of quartz plates. (*From Raymond A. Heising, "Quartz Crystals," D. Van Nostrand Company, Inc., Princeton, N.J., 1946.*)

of a piece of crystal depends on its dimensions as well as on how it is cut with respect to the different faces. The way in which the resonant frequency varies with temperature also depends on the cut.

The performance characteristics of a crystal also depend on the particular mode of vibration. There are four basic modes of mechanical vibration, of which three are in commercial use. These are the flexure, shear, and longitudinal or extensional modes. The fourth which is seldom used is the torsional mode. Crystals operating in the thickness-shear mode can be made to operate on odd overtones. Such operation may be referred to as *third-mode* or *fifth-mode* operation. Seventh- and higher-overtone oscillation is possible but is seldom used commercially. Crystal

cuts are divided into two groups which are described as *rotated X-cut* and *rotated Y-cut* crystals. Figure 2-4 illustrates the type of vibration, and Fig. 2-5 shows the approximate locations of the various cuts in a crystal. The most widely used cut is the AT. This cut has a frequency factor (frequency in megacycles times thickness in thousandths of an inch) of 66.4, which makes it nearly the thinnest band for its frequency of any

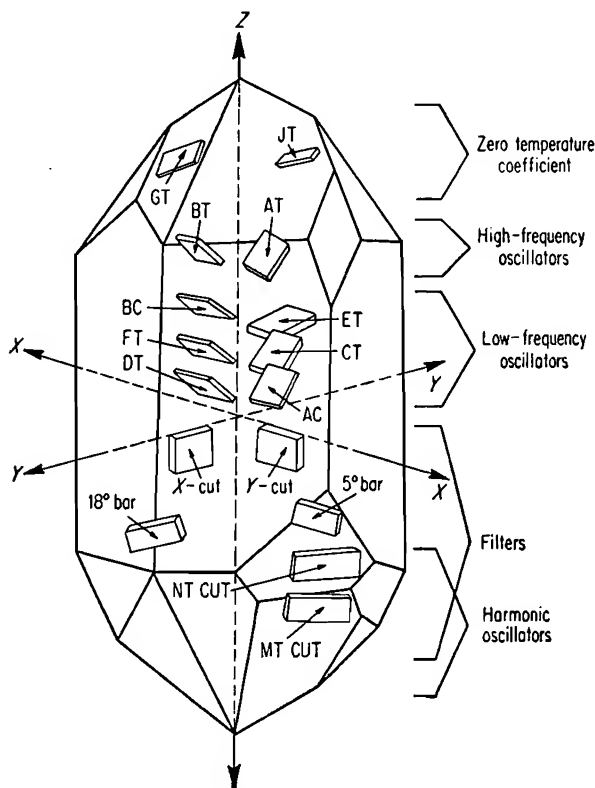


FIG. 2-5. Common crystal cuts. (Standard Crystal Co.)

cut. The GT cut has the best temperature coefficient. This cut is a good example of the fact that the associated circuit cannot always be chosen without regard to the cut. It is intended to operate in the width mode, but, if it is plugged into an untuned oscillator such as the Pierce, it may oscillate on the length mode which will be about 14 per cent lower in frequency than desired. Furthermore, the temperature coefficient is much poorer in the length mode.

The lower-frequency X cuts are more widely used for crystal filter work than for crystal oscillators.

2-4. Crystal Mountings. Crystal plates are generally mounted in *crystal holders* which refer to the mounting and housing assemblies. A

crystal unit refers to the assembly of holder and crystal. The crystal unit may in turn be mounted in a *crystal oven* for maintaining a constant temperature.

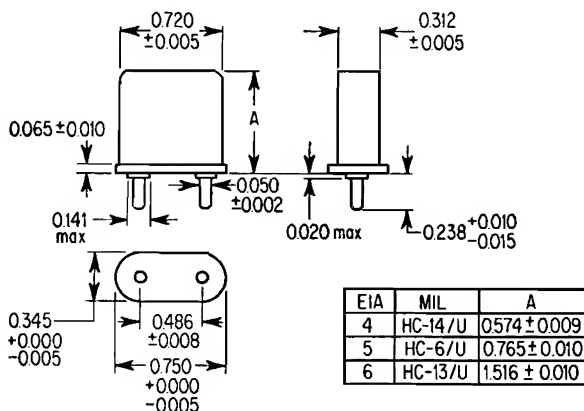


FIG. 2-6. Dimensions of EIA crystal holders 4, 5, and 6. (EIA.)

There are many methods of mounting crystals, but the two in most general use are the *wire* and *pressure* mountings. The wire-mounted crystals are cemented to supporting wires which provide a firm but flexible mounting to protect the crystal against external vibration and shock. The wire-mounted crystal is housed in a sealed glass or metal container. Most sealed units are sealed in dry air; but for optimum performance of certain types such as those operating in the flexure mode, the housing is evacuated. The damping effect of the air is avoided in such a housing. Pressure-mounted crystals are generally housed in plastic holders which are sealed with neoprene gaskets. The Electronic Industries Association has standardized a number of crystal holders in their standard RS-192 (14), entitled "Definitions and Dimensional Characteristics of Quartz Crystal Units." The dimensions of the most commonly used holders are given in Figs. 2-6 and 2-7. Most crystal manufacturers make a practice of identifying the complete unit which includes both the holder and the crystal.

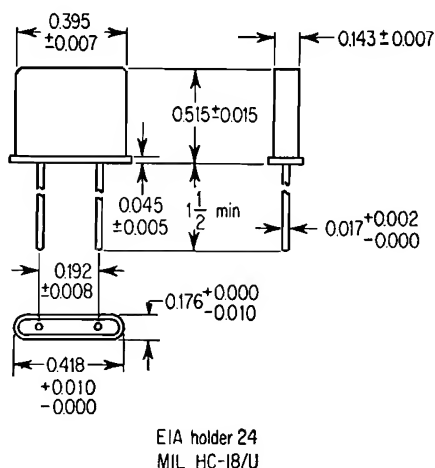


FIG. 2-7. Dimensions of EIA crystal holder 24. (EIA.)

2-5. Crystal Ovens. Where high stability is required, the temperature of the air surrounding the crystal unit is maintained at some value higher than the highest ambient temperature expected. As crystal-grinding techniques have advanced, it has been possible to control the temperature coefficient with a consequent reduction in the required accuracy of temperature control for a given stability. It is possible to cut crystal plates in such a way that the temperature coefficient can be made to change from positive to negative at a selected temperature. The temperature at which the temperature coefficient goes through zero is referred to as the *turning point*, and to obtain best stability the turning point will be placed at the operating temperature of the oven. The type of crystal oven most commonly encountered is the small plug-in type of unit intended for mounting in standard tube sockets. For fixed station transmitters or for frequency standards, the crystal ovens may be large compartments housing the entire oscillator circuit (15). Such boxes may employ thermistor-bridge thermostats with electronic amplification. The small plug-in ovens generally are controlled by bimetallic thermostats, which turn the heater power on or off directly. A front-panel indicator lamp is usually connected across the crystal heater, and the constant flashing of the lamp indicates normal operation.

2-6. Crystal-frequency Correlation. The frequency at which a crystal-controlled oscillator operates depends not only upon the physical dimensions of the crystal but on the associated circuit. At one time it was the usual practice for equipment manufacturers to supply the crystal manufacturers not only with the circuit diagram but often with the actual oscillator section of the equipment as well. This procedure had many disadvantages, and a system for crystal-frequency correlation using standard test sets and MIL or EIA specifications was developed. For example, a military type CR-18/U crystal unit operates within 0.005 per cent of its nameplate frequency when used in a parallel-resonant circuit with a load capacitance of 32 μf . If the same crystal were to be used in a circuit which permitted the crystal to oscillate at its series-resonant frequency, the oscillator output frequency would be lower. A type CR-19/U crystal unit would be properly specified for the series-resonant circuit.

2-7. Military Standard Crystal Units. A list of the military crystal specifications from specification MIL-C-3098C (16) is given in Table 2-1. The categories shown refer to the permissible uses assigned for military equipment as of 1959 (17). The categories are as follows:

1. Recommended for use in new equipment
2. Recently developed and not yet standardized
3. Allowed for replacement purposes
4. Used in past but no longer considered suitable

TABLE 2-1. MILITARY STANDARD CRYSTAL UNITS*

Crystal type	MIL holder	Frequency range, kc	Frequency tolerance, %	Temperature range, °C	Resonance	Load capacitance, μmf	Mode	Category
CR-1A/AR	CR-1A	3,600-15,000	± 0.02	-55 to + 90	Parallel	35	Fundamental	3
CR-2/U	HC-17/U	200	± 0.009	-40 to + 70	Parallel	125	Fundamental	3
CR-2/U	HC-17/U	500	+0.004 -0.013	-40 to + 70	Parallel	64	Fundamental	3
CR-3/U	HC-17/U	300-600	± 0.02	-40 to + 70	Parallel	...	Fundamental	3
CR-4/U	HC-17/U	500-1,200	± 0.03	-40 to + 70	Parallel	...	Fundamental	3
CR-5/U	FT-243	2,000-10,000	± 0.02	-55 to + 90	Parallel	25	Fundamental	3
CR-6/U	FT-243	2,000-10,000	± 0.02	-40 to + 70	Parallel	13	Fundamental	3
CR-7/U	Not used							4
CR-8/U	FT-243	1,000-10,000	± 0.02	-40 to + 70	Series	...	Fundamental	3
CR-9/U	Replaced by CR-24/U							4
CR-10/U	FT-243	5,000	± 0.005	-40 to + 70	Parallel	25		3
CR-11/U	Western Electric	1.8-50						4
CR-12/U	Not used							4
CR-13/U	FT-243	455-5,250	± 0.02	-40 to + 70	Series	...	Fundamental	3
CR-14/U	Not used							4
CR-15B/U	HC-21/U	80-200	± 0.01	-40 to + 70	Parallel	32	Fundamental	1
CR-16B/U	HC-21/U	80-200	± 0.01	-40 to + 70	Series	...	Fundamental	1
CR-17/U	Replace by CR-24/U							4
CR-18A/U	HC-6/U	800-20,000	± 0.005	-55 to +105	Parallel	32	Fundamental	1
CR-19A/U	HC-6/U	800-20,000	± 0.005	-55 to +105	Series	...	Fundamental	1
CR-20/U	11-pin base	422.5	75	No longer specified			4
CR-21/U	Not used							
CR-22/U	Not used							
CR-23/U	Replace by CR-51U, CR-52/U, CR-53/U, or CR-54/U							4
CR-24/U	HC-10/U	15,000-50,000	± 0.005	-55 to +105	Series	...	3rd (15-25) 5th (25-50)	1
CR-25A/U	HC-6/U	200-500	± 0.01	-40 to + 70	Series	...	Fundamental	1
CR-26A/U	HC-6/U	200-500	± 0.002	+70 to + 80	Series	...	Fundamental	1
CR-27A/U	HC-6/U	800-20,000	± 0.002	+70 to + 80	Parallel	32	Fundamental	1
CR-28A/U	HC-6/U	800-20,000	± 0.002	+70 to + 80	Series	...	Fundamental	1
CR-29A/U	HC-21/U	80-200	± 0.002	+70 to + 80	Parallel	32	Fundamental	1
CR-30A/U	HC-21/U	80-200	± 0.002	+70 to + 80	Series	...	Fundamental	1
CR-31/U	Not used							4
CR-32A/U	HC-6/U	10,000-75,000	± 0.002	+70 to + 80	Series	...	3rd (10-52) 5th (52-75)	1
CR-33A/U	HC-6/U	10,000-25,000	± 0.005	-55 to +105	Parallel	32	3rd	2

* Courtesy of Armed Services Electro Standards Agency.

TABLE 2-1. MILITARY STANDARD CRYSTAL UNITS* (Continued)

Crystal type	MIL holder	Frequency range, kc	Frequency tolerance, %	Temperature range, °C	Resonance	Load capacitance, μf	Mode	Category
CR-34/U	Not used							4
CR-35A/U	HC-6/U	800-20,000	± 0.002	+80 to + 90	Series	...	Fundamental	1
CR-36A/U	HC-6/U	800-20,000	± 0.002	+80 to + 90	Parallel	32	Fundamental	1
CR-37A/U	HC-13/U	90-250	± 0.02	-40 to + 70	Parallel	20	Fundamental	1
CR-38A/U	HC-13/U	16-100	± 0.012	-40 to + 70	Parallel	20	Fundamental	1
CR-39/U	HC-15/U	160-330	± 0.003	+25	Series	...	Fundamental	3
CR-40/U	HC-15/U	160-330	± 0.003	+65 to + 75	Series	...	Fundamental	3
CR-41/U	Not used							
CR-42A/U	HC-13/U	90-250	± 0.003	+70 to + 80	Parallel	32	Fundamental	1
CR-43/U	HC-16/U	80.860	± 0.01	+25	Parallel	45	Fundamental	2
CR-44/U	HC-6/U	15,000-20,000	± 0.002	+80 to + 90	Parallel	32	Fundamental	3
CR-45/U	HC-6/U	455	± 0.02	-40 to + 70	Series	...	Fundamental	1
CR-46A/U	HC-6/U	200-500	± 0.01	-40 to + 70	Parallel	20	Fundamental	1
CR-47A/U	HC-6/U	200-500	± 0.002	+70 to + 80	Parallel	20	Fundamental	1
CR-48/U	Replace by CR-18/U							
CR-49/U	Replace by CR-38/U							
CR-50A/U	HC-13/U	16-100	± 0.012	-40 to + 70	Series	...	Fundamental	1
CR-51A/U†	HC-6/U	10,000-61,000	± 0.005	-55 to +105	Series	...	3rd	1
CR-52A/U	HC-6/U	10,000-61,000	± 0.005	-55 to +105	Series	...	3rd	1
CR-53A/U†	HC-6/U	50,000-87,000	± 0.005	-55 to +105	Series	...	5th	1
CR-54A/U	HC-6/U	50,000-125,000	± 0.005	-55 to +105	Series	...	5th	1
CR-55/U	HC-18/U	17,000-61,000	± 0.005	-55 to +105	Series	...	3rd	1
CR-56A/U	HC-18/U	50,000-125,000	± 0.005	-55 to +105	Series	...	5th	1
CR-57/U	HC-6/U	500	± 0.001	+80 to + 90	Parallel	32	Fundamental	1
CR-58A/U	HC-17/U	800-20,000	± 0.005	-55 to +105	Parallel	32	Fundamental	2
CR-59A/U	HC-18/U	50,000-125,000	± 0.002	+80 to + 90	Series	...	5th	1
CR-60A/U	HC-18/U	5,000-20,000	± 0.005	-55 to +105	Series	...	Fundamental	1
CR-61/U	HC-18/U	17,000-61,000	± 0.002	+80 to + 90	Series	...	3rd	1
CR-62/U	HC-6/U	800-850	± 0.001	+70 to + 80	Parallel	32	Fundamental	1
CR-63A/U	HC-6/U	200-500	± 0.01	-40 to + 70	Parallel	20	Fundamental	1
CR-64/U	HC-18/U	7,000-20,000	± 0.005	-55 to +105	Parallel	30	Fundamental	1
CR-65/U	HC-6/U	10,000-61,000	± 0.001	+70 to + 80	Series	...	3rd	1

* Courtesy of Armed Services Electro Standards Agency.

† These are pressure-mounted units, and a drive level of 20 mw is allowed.

2-8. Crystal-oscillator Circuits. Crystal oscillators may be classified as being crystal-controlled or crystal-stabilized. The latter type will oscillate if the crystal is removed and is not in general use in transmitters because it may not be obvious to the operator that the output frequency is not being controlled by the crystal.

Most oscillator circuits can be classified as being of the *series-resonant* or *parallel-resonant* type. The parallel-resonant type circuit lends itself to small frequency adjustment by means of a variable shunt capacitor. The series-resonant circuits can provide greater frequency stability and are also used when it is desired to operate a crystal on a harmonic or *overtone*.

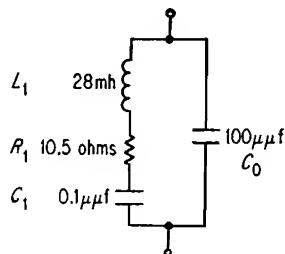


FIG. 2-8. Typical crystal equivalent circuit. (IEEE.)

The simplified equivalent circuit of a crystal plate is given in Fig. 2-8 (18). Typical circuit values are given for a 3-Mc AT-cut plate.

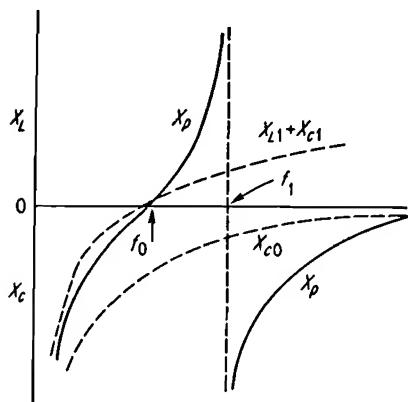


FIG. 2-9. Crystal-reactance characteristic.

$$\begin{aligned}
 X_p &= \frac{(X_{L1} + X_{C1})X_{C0}}{X_{L1} + X_{C1} + X_{C0}} \\
 &= 0 \text{ when } X_{L1} = -X_{C1} \quad (\text{series-resonant}) \\
 &= \infty \text{ when } X_{L1} + X_{C1} = -X_{C0} \quad (\text{parallel-resonant})
 \end{aligned}$$

of X_{C0} will be reduced for any given frequency, and the lower dotted line will effectively move up. It will be seen that f_1 will become a new, lower value. The minimum frequency at which a crystal can be made to oscillate will be the series-resonant frequency.

The shunt capacity of the crystal itself depends on the cut and mode of vibration. Typical ratios between C_0 and C_1 are 150 for an X-cut crystal and 1,000 for an AT-cut crystal.

The Q in the case shown is about 50,000. The way in which the reactance of the crystal and of the two branches varies with frequency is shown in Fig. 2-9. f_0 is the series-resonant frequency and is the frequency at which $X_{C1} = X_{L1}$. f_1 is the parallel-resonant frequency, and this is the frequency at which the reactance of C_0 is equal to the net reactance of the L_1C_1 branch. The solid lines represent the reactance of the crystal plus any external capacitance plotted against frequency. The dotted lines represent the reactances of the individual branches.

If the value of C_0 (which includes internal effective capacity of the crystal itself plus any external shunting capacity) is increased, the value

2-9. Parallel-resonant Circuits. Most parallel-resonant circuits may be classified as either Miller or Pierce circuits. One exception is a crystal-controlled multivibrator circuit. Basically the Miller circuit is a Hartley oscillator with the crystal replacing the grid-to-cathode inductance. The Pierce circuit is equivalent to a Colpitts oscillator with the crystal replacing the plate-to-grid inductance. The position of the ground has nothing to do with the classification of the circuit, and any one of the tube electrodes may be bypassed to ground in either type of circuit. The Pierce oscillator does not require a tuned circuit. A resonant circuit is required in the Miller oscillator, and the frequency depends somewhat on this circuit; this type of circuit is therefore less stable than the Pierce circuit. On the other hand, the Miller circuits are capable of much greater output and for this reason are probably in

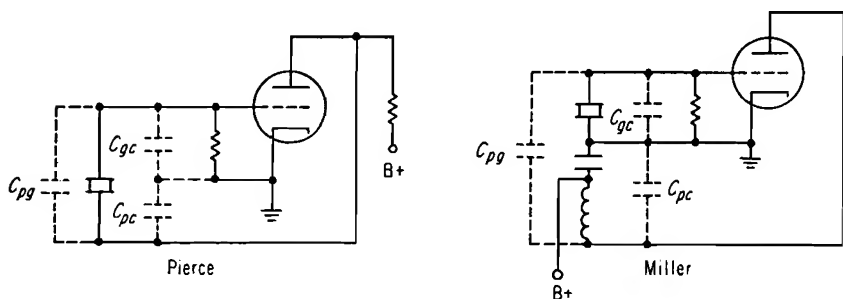


FIG. 2-10. Crystal oscillator circuits.

greater use than the Pierce circuits. When economy or size is deemed more important than stability, the Miller circuit is more useful because it usually requires one less stage of amplification.

There are a number of variations of both the Miller and Pierce circuits. In Ref. 13 there are described 43 Pierce circuits and 31 Miller circuits. Typical Pierce and Miller circuits are given in Fig. 2-10 (19,20). When correlated crystals, such as those made to military or EIA specifications, are used, the circuit capacities which load the crystal must be of specified value. This is usually $32\ \mu\text{f}$ in the range above 500 kc and $20\ \mu\text{f}$ below 500 kc.

2-10. Series-resonant Circuits. In any series-resonant circuit, the crystal must be connected between points which will be at a low impedance with respect to each other when the circuit is operating. Most series-resonant circuits will oscillate when the crystal is replaced by a short circuit. There are a number of series-resonant circuits, the one probably in widest use being the Butler circuit. Another group of series circuits are the bridge-stabilized oscillators. These are capable of very

high stabilities and for certain applications, such as single sideband, may be found in transmitters.

The basic Butler oscillator is shown in Fig. 2-11. The plate circuit may be tuned to an odd harmonic of the fundamental crystal frequency, and the crystal will then operate in an overtone mode. It is important to note that the frequency of the overtone operation may not be a precise multiple of the fundamental frequency, and such circuits should be used

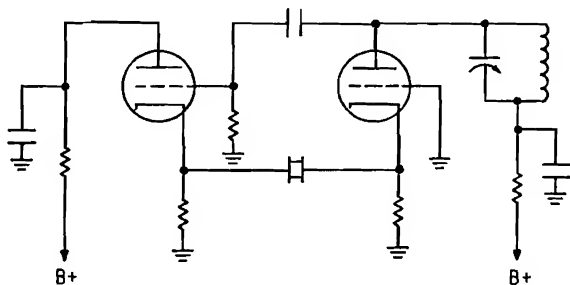


FIG. 2-11. Butler circuit.

only with crystals which have been ground and checked specifically at the output frequency to be used. There are a number of variations (21) of the Butler circuit including single-triode types which use tapped inductances to achieve the correct phase relationships (13). Crystal oscillators making use of transistors are frequently of the series-resonant type. A circuit making use of a silicon mesa transistor operating at 70 Mc is shown in Fig. 2-12 (22). The junction-type transistors are

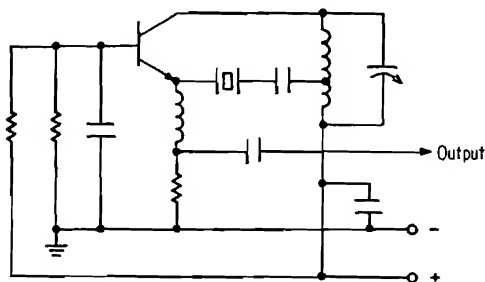


FIG. 2-12. 70-Mc transistor oscillator. (Courtesy of Texas Instruments, Inc.)

suitable for lower-frequency applications, and a 100-ke circuit is shown in Fig. 2-13 (23).

2-11. Drive Level. Crystal plates may be cracked or may have their resonant frequencies shifted if an excessive drive level is applied. The drive level can be determined when the voltage across the crystal and its effective series resistance is known. Allowable drive levels vary between 0.1 and 20 mw (milliwatts). For example, a type-CR-18/U crystal has a

resonant resistance of about 175 ohms at 3,000 kc and is limited to a drive level of 5 mw.

The drive level will also depend on the feedback in the oscillator circuit. This is the plate-to-grid capacity in the Miller circuit and ratio of C_g to C_p in the Pierce oscillator. The drive level will be reduced by reducing the plate-to-grid capacity or increasing the C_g/C_p ratio. The drive level also depends on a property of the crystal itself which is usually called its *activity*. The activity of a crystal (which can be roughly defined as its ability to oscillate) is greater when its series-equivalent resistance is less. It should be noted that the allowable drive level for a crystal unit is proportional to its size, and the small wire-mounted units can tolerate much less drive than the older larger types.

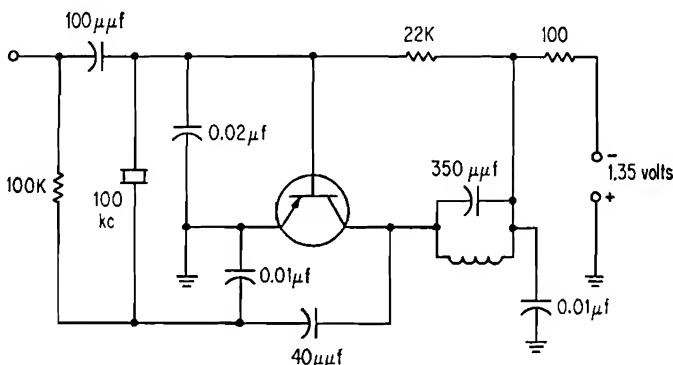


FIG. 2-13. Junction transistor crystal oscillator. (*Electronic Design.*)

2-12. Variable-frequency Crystal Oscillators. In most military and commercial applications, crystal oscillators are operated at fixed frequencies, and operation within the specified frequency tolerance is achieved by maintaining a close grinding tolerance and by temperature control where necessary. However, there are applications where operation on a certain exact frequency is desired. In other applications a controllable variation is required such as for frequency-shift keying, automatic-frequency-control circuits, and phase-locking circuits in diversity systems. The frequency of a parallel-resonant type of crystal oscillator can be varied slightly by connecting a small variable capacitor either in series or in shunt with the crystal plate. The variation can be increased by as much as four times by associating an inductance with the capacity in either the series or the parallel case. The attainable variation varies with the type of cut used. Variations of 0.1 per cent are not difficult to achieve with AT-cut plates when inductance-capacity combinations are used. A variable-frequency crystal-oscillator circuit is shown in Fig. 2-14. The inductance across the crystal is fixed and has a

value of 35 to 70 μh for a frequency of 3 Mc. The variable capacitor has a maximum value of 25 μmf , and a frequency variation of 4 to 7 kc should be attainable (18). There will usually be some sacrifice in stability when such circuits are used. Since X-cut crystals have a lower inherent shunt capacity, it is possible to shift their frequencies by a greater amount. A limitation of this type of circuit is that spurious responses may be encouraged in some crystals. Complicated couplings to other modes of oscillation exist in AT- and BT-cut crystals, and the tendency for the crystal frequency to jump to another mode increases as the shunt inductance is increased. For circuits where fixed-frequency operation is required, the shunt inductance should not be used, and the crystal should be ground so that the frequency will lie within the limits of variation provided by a small shunt capacity.

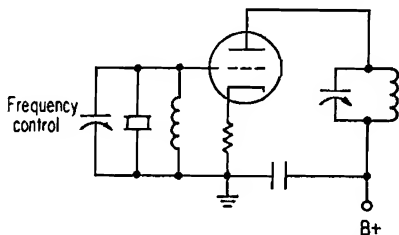


FIG. 2-14. Variable-frequency crystal oscillator.

2-13. Frequency Stability of Crystal Oscillators. When discussing frequency stability, the time over which observations are made must be stated to qualify the figure. *Short-term* stability may refer to variations over 1 hr or 1 day, whereas *long-term* stability may refer to a period of several months. The short-term stability will be governed by:

1. Temperature variation of crystal and associated circuit components
2. Effect of supply-voltage changes
3. Adjustment of oscillator circuit controls or following circuits

The long-term stability is mostly governed by aging of the crystal plate and by long-term variations in the characteristics of the tube, transistor, or associated components. The aging of crystal plates is apparently due to a gradual removal of minute particles of pulverized quartz on the plates. This occurs even with acid-etched plates which have been carefully washed. Most of the aging variation will take place during the first 14 days of operation. The other causes of instability can all be more or less removed by close temperature control and isolation of the crystal from the circuit as much as possible. In certain transmitter applications such as the control for loran¹ equipment, a very high degree of frequency stability is necessary. A stability of a few parts in 10^{10} per day can be achieved by using GT-cut crystals associated with suitable circuitry such as a high-sensitivity bridge-balancing frequency-correction system as described by Sulzer (24-26). Simpler methods of improving oscillator stability include dilution of tube input capacity variations by

¹ Long-range navigation.

inserting a series capacitor between the crystal and the tube and connecting a larger capacity across the tube input. In this case values would be chosen to maintain the correct load capacity across the crystal. The effects of circuits following the crystal stage can be made negligibly small by the use of electron coupling such as in the circuit given in Fig. 2-15.

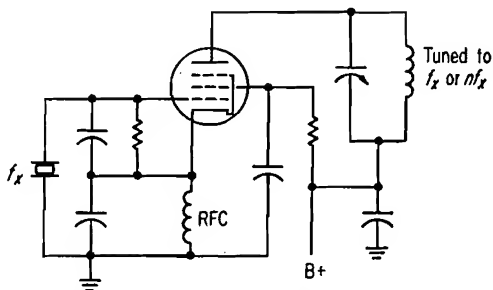


FIG. 2-15. Electron-coupled Pierce oscillator.

In this circuit the plate circuit may be tuned to the fundamental or to a harmonic of the crystal frequency, and in any case the adjustment of the plate circuit will have a negligible effect on the frequency.

2-14. Frequency Control in F-M Transmitters. In an f-m transmitter the frequency is deviated by an amount which is proportional to the amplitude of the modulating signal and swings at a rate corresponding to the modulating frequency. At the same time, government regulations

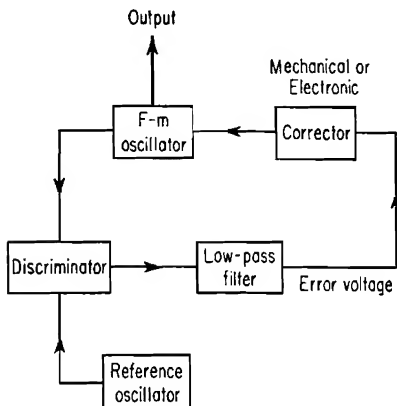


FIG. 2-16. Center-frequency stabilization of f-m oscillator.

require that the average frequency be maintained within limits which are much smaller than the total variation. In equipment where phase modulation is used, the problem is simple because a crystal-controlled oscillator can be phase-modulated. The average frequency of frequency-modulated oscillators is held constant by variations of the basic system shown in Fig. 2-16. A crystal-controlled signal is compared with the average frequency of the modulated oscillator or a subharmonic of it by

a phase or frequency comparator which generates an error voltage. This error voltage may control a motor-driven capacitor which corrects the frequency or may motor-drive an adjustable klystron in the case of an oscillator in the kMc region. The error voltage may also control the bias of the same reactance tube that modulates the oscillator. When a

phase comparator is used, the modulated oscillator will be *locked* to the crystal oscillator, and the frequencies of the compared signals will be identical even though there is a phase difference and an error voltage is being generated.

When a frequency comparator is used, the frequency at which the discriminator generates a zero error voltage will depend on the adjustment of the discriminator itself as well as on the frequency of the reference oscillator. For a system requiring high stability, it would be necessary to temperature-control the discriminator or at least use temperature-compensated components. This precaution is not necessary with a

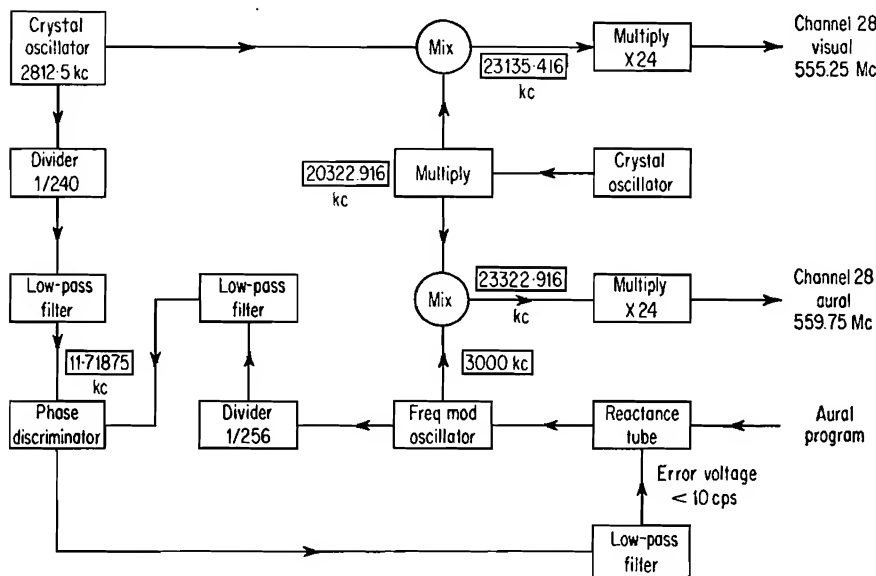


FIG. 2-17. TV-transmitter frequency control. (AIEE.)

phase discriminator, but there is one important limiting factor to be considered when a phase discriminator is used. This is that the phase modulation which appears on the signal which is applied to the phase discriminator should not be allowed to go much over $\pm 90^\circ$, or the system will not remain in lock. This difficulty is avoided by dividing the f-m signal down to a point where the phase modulation is less than 90° . The relation between phase and frequency modulation will be discussed in more detail in the chapter on modulation. An example of frequency-control circuits of this type for an f-m transmitter is given in Fig. 2-17 (27,28).

2-15. Frequency Control of Microwave Oscillators. For transmitters operating above 10,000 Mc, it is not practical to multiply up from a stable low-frequency crystal oscillator. The usual practice in these ranges is to

use self-excited oscillators such as reflex klystrons. Fortunately the oscillating frequency of these devices can be varied slightly by adjustment of the repeller or reflector voltage. The klystron output frequency can then be compared with some reference, and an error voltage generated and utilized to correct the klystron frequency.

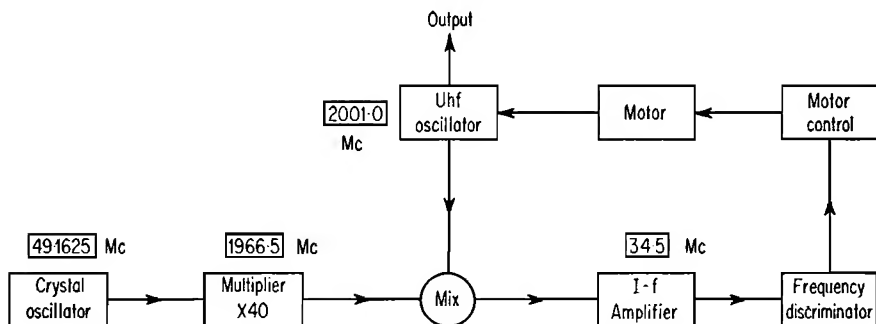


FIG. 2-18. Microwave oscillator control.

The comparison can be effected in a number of ways. Since the reference signal can be at a low voltage level, it is now possible to use direct-frequency multiplication from a crystal-controlled source. The last multiplication may be as much as 40 times and can be accomplished with silicon diode crystals of the microwave mixer type. A block diagram of

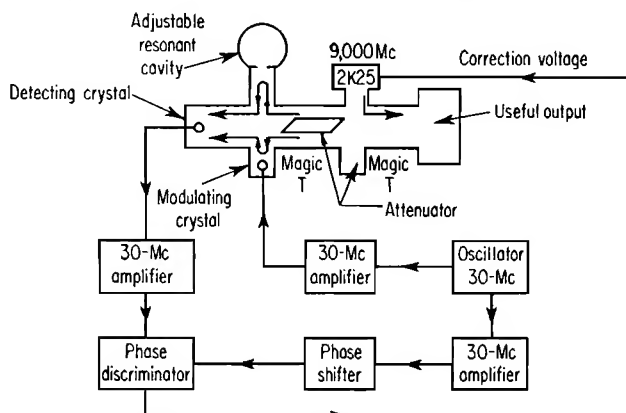


FIG. 2-19. Microwave oscillator stabilizer. (*Proc. IRE.*)

this system is given in Fig. 2-18. It is possible to build microwave cavities which have much higher Q 's and are more stable than the cavities used in the klystrons. Such cavities can be accurately calibrated and used as discriminators at the microwave frequency. Such a discriminator would generate an error voltage which could be used to control the klystron repeller voltage. Another method is shown in Fig. 2-19. This

has been referred to as an *i-f stabilizer* (29). The operating frequency is determined by the reference cavity. When the oscillator frequency is not the same as the cavity frequency, a 30-Mc signal will appear at the detector crystal, and this signal is amplified to generate a correction voltage which brings the oscillator to the correct frequency. When the oscillator frequency is the same as the resonant frequency of the cavity, no *i-f* voltage is generated.

2-16. Frequency Synthesizers. A frequency synthesizer is a device which will generate a large number of discrete stable frequencies while using a small number of crystals, in some cases only one. A large number of different types of synthesizers have been built for both military and commercial applications. A few will be described in some detail to illustrate the principles involved (30). Synthesis systems may be divided into two broad categories:

1. *Direct synthesizers.* In this type, the signal is built up by division, multiplication, addition, or subtraction of components derived from reference crystal oscillators.

2. *Controlled oscillators.* In this type, the required frequency range is covered by a variable oscillator, and this is "disciplined" or corrected to an exact stable frequency by comparison with signals derived from the reference source.

The second type has the slight advantage that if all the synthesizer circuits fail, the transmitter may still be driven by the variable-frequency oscillator. For ssb systems, however, the transmitter could probably not be used with an unstabilized variable-frequency oscillator. On the other hand, in military mobile equipment it is desirable to have the emergency facility available.

Elements of Synthesizers. A good discussion of frequency synthesizers may be found in Buchanan (31). The basic circuit elements required to build a synthesizer include

1. Reference oscillators
2. Frequency multipliers or dividers
3. Mixers (addition or subtraction)
4. Filters

The reference oscillators are usually crystal-controlled although a 1,000-cps tuning-fork oscillator was used in one synthesizer (32).

2-17. Frequency Multipliers and Dividers. Quartz crystals can be made to operate on overtones up to 85 Mc or more. It may be more convenient, however, to start with a much lower crystal oscillator and to obtain the desired higher frequencies by multiplication. Harmonics can be generated by passing a signal through a nonlinear circuit element

such as a diode or amplifier which is biased beyond cutoff for at least part of the driving cycle. The desired harmonic frequency is then selected with a tuned circuit. (See Sec. 3-18 for a discussion of multipliers.)

It may also be necessary to generate frequencies which are lower than the reference-oscillator output, and this is done in frequency dividers. Multivibrators, phantastrons, and blocking oscillators are typical frequency-dividing circuits (33). The *bistable* multivibrator is a common frequency-dividing circuit. This circuit has two stable states and basically has two triodes or transistors. One section will be conducting while the other will be cut off. An input pulse will reverse the condition, and the number of output pulses that appears at each section will be exactly one-half the number of input pulses.

2-18. Mixers. Mixers are used in synthesizers to generate a desired stable frequency which will be the sum or difference of available stable frequencies. The advantage of the mixing process is that a high frequency can be derived from variable-frequency oscillator that is free-running and relatively unstable without multiplying the instability. For example, suppose that a 1-Mc variable-frequency oscillator is stable within 50 cps over a given period. If this is multiplied up to 30 Mc, the frequency variation would be 1,500 cps; but if it is mixed with a 29-Mc crystal oscillator that has a stability of 30 cps, the stability of the mixed output can be no worse than $50 + 30$ or 80 cps.

When two signals are applied to a mixer, the output will contain these same two frequencies, the sum and difference of the two frequencies, and also sums and differences of the harmonics of the input frequencies (34). Filters must follow the mixer to select the desired frequency.

An important problem that comes up is that the *spurious* frequency (sums or differences of harmonics) may be so close to the desired frequency that the filter cannot separate out the desired frequency unless it is made extremely narrow. For example, if 1,000 kc is mixed with 501 kc, there will be an output of 1,501 kc, but there will also be an output at 3×501 or 1,503 kc. The 1,501- and 1,503-kc signals will beat together and generate a 2-kc tone.

When the spurious output coincides with the desired sum or difference, the spurious output is said to be the result of a *crossover*. The most practical method of avoiding this difficulty is to select input frequencies that will not result in crossovers. It is important also that the crossover will be far enough from the operating point so that the filters can attenuate them sufficiently.

A crossover chart is given in Fig. 2-20 (35). f_0 and f_x are the frequencies to be mixed, f_0 being the lower. The unwanted spurious output is f_s . The horizontal scale is the ratio f_0/f_x , and the vertical scale is the ratio f_s/f_x . The upper heavy slant line is the operating point when the

desired output is $f_x + f_0$, and the lower heavy line is the operating point when the desired output is $f_x - f_0$.

To use the chart, determine the ratio between the input frequencies which have been selected (f_0/f_x) and trace upward from this value until the heavy difference line or the heavy sum line is reached, depending on whether the difference or sum is desired. From the operating point on the heavy lines, trace up or down to the nearest slant line, and then read the value of f_s/f_x from the scale at the left edge of the chart. The value of the spurious output frequency can then be determined.

For example, suppose that it is desired to generate 27 Mc. This can be done by feeding 30 and 3 Mc into a mixer and using the difference. $\frac{3}{30} = 0.1$, and the nearest slant line to the point where 0.1 crosses the heavy difference line is the $f_x - 2f_0$ line. The value of f_s/f_x , which is directly below the operating point on the harmonic line, is 0.8. This means that the spurious frequency is 24 Mc, and this can be easily separated from the desired 27 Mc in a filter following the mixer. The chart will give a rough indication of the spurious frequency, and the exact value can be determined from the designation of the harmonic line. In the above example the spurious was due to $f_x - 2f_0$ or

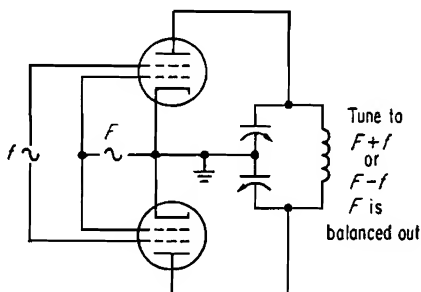
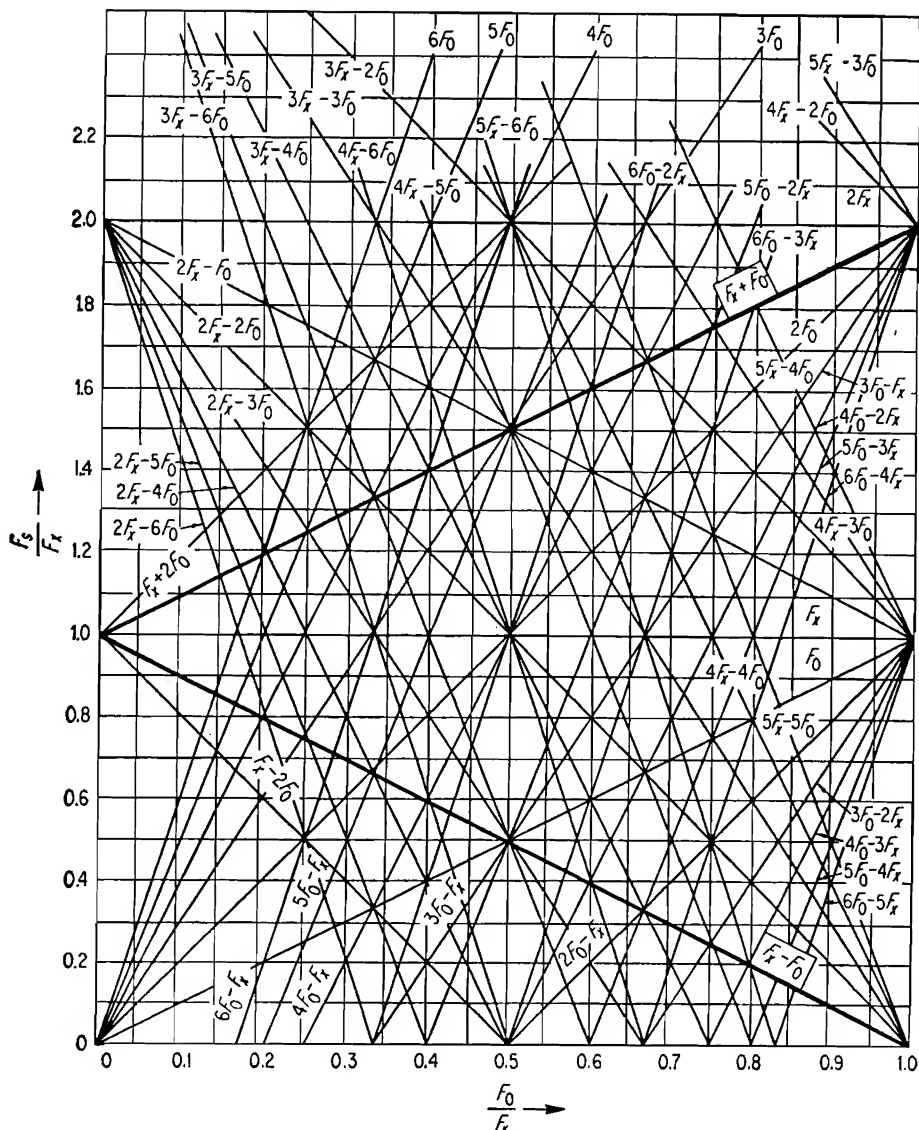


FIG. 2-20. Simplified balanced-modulator circuit.

$$30 - 2 \times 3 = 24 \text{ Mc}$$

When a wide tuning range is to be covered, one of the input frequencies to the mixer will vary over this range. Spurious outputs can best be avoided by choosing a ratio of at least 8:1 between the two input frequencies. If the ratio becomes too high, it becomes difficult to separate the desired output from the higher of the input frequencies.

2-19. Filters. The filters following the mixer may be fixed or tunable, depending on the ranges to be covered and the location of the spurious outputs. It is general practice to reduce spurious outputs from a synthesizer to values which are at least 80 db below the desired output. The filter design can sometimes be simplified by making use of a *balanced-modulator* type of mixer. This type of mixer will greatly attenuate the input which is applied in push-pull to the modulator grids. The filter design is simplified because the closest frequency that must be attenuated is the unwanted sum or difference of the two input frequencies. Balanced modulators are discussed in detail in Sec. 6-16.

FIG. 2-21. Mixer harmonic chart. (*Electronics.*)

2-20. Drift-canceled Oscillator. A circuit which is frequently used in conjunction with synthesizers is the drift-canceling circuit (30). In a typical example of a generator producing outputs 10 kc apart, a variable-frequency oscillator which operates in the range 600 to 700 kc is first mixed with the output of a spectrum generator which has signals every

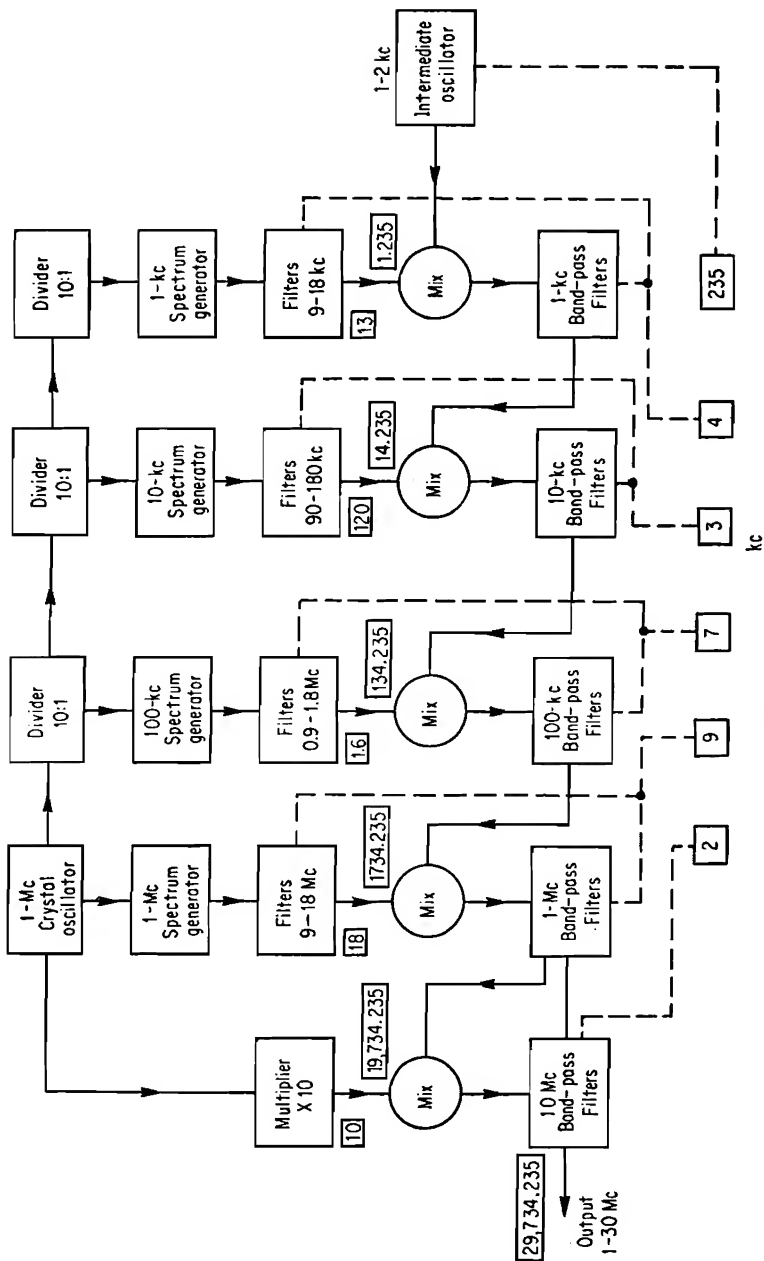


Fig. 2-22. Direct synthesizer.

10 kc from 350 to 450 kc. The output of the first mixer is connected to a filter which has a bandwidth of only 6 kc from 247 to 253 kc. The filter output is mixed again with the variable-frequency oscillator to produce a single signal in the 350- to 450-kc range. Consider, for example, that the variable-frequency oscillator is set at 640 kc. This will heterodyne with the 390-kc signal from the spectrum generator, and the mixer output will be 250 kc. In the second mixer, 250 kc is subtracted from 640 to obtain 390 kc. Suppose, now, that the variable-frequency oscillator drifts to 641 kc. The first mixer output will be 251 kc, but the

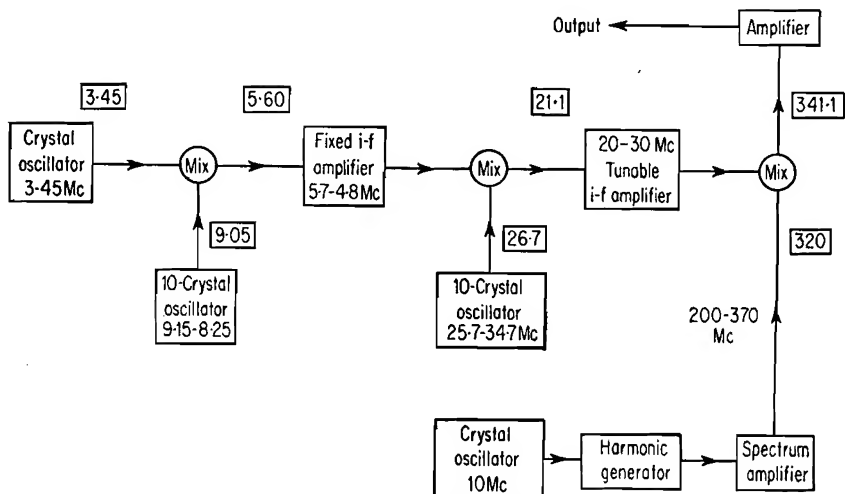


FIG. 2-23. Multicrystal synthesizer.

output from the second mixer will still be exactly 390 kc. If the variable-frequency oscillator drifts to 645 kc, there will be output signals from the first mixer at 245 and 255 kc; but these will not pass through the filter, and there will be no output from the second mixer. As the variable-frequency oscillator is continuously tuned from 600 to 700 kc, there will be a succession of output signals from the second mixer; these signals are exactly at 10-kc intervals from 350 to 450 kc. The level of the spurious output signals will depend on the characteristics of the filter.

2-21. Synthesizing Systems. Early synthesizers were developed for frequency-measuring purposes and described by Finden (36) of the Plessey Company, England and by Petersen (37) of RCA. In these units the basic reference oscillator is operated at 1 Mc. Stabilized outputs were generated at 1-kc intervals up to 30 Mc. The reference oscillator was divided or multiplied, and the resulting frequencies were mixed and filtered to obtain the desired outputs. Figure 2-22 shows the Petersen system.

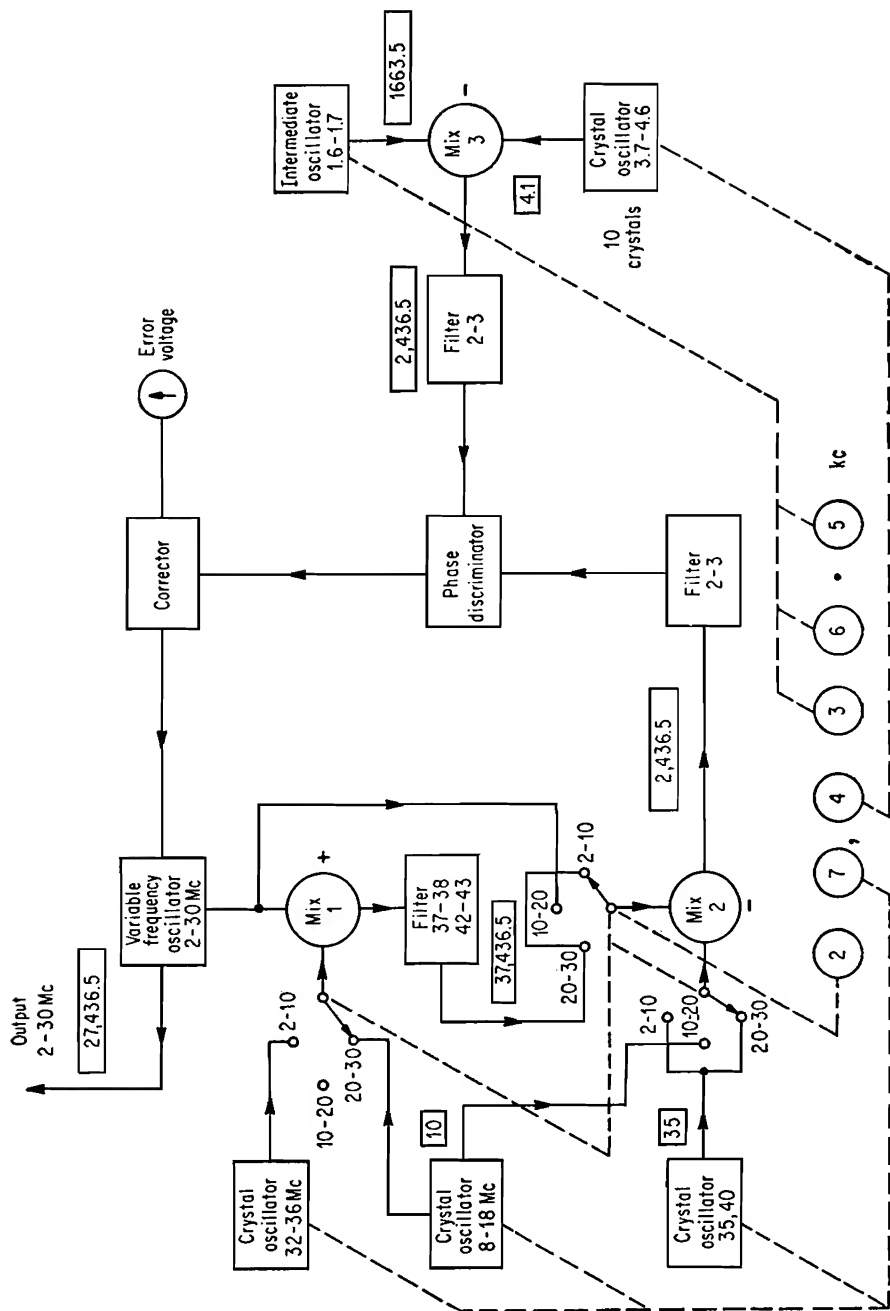


Fig. 2-24. Phase-locked variable-frequency oscillator (mobile) synthesizer. (ITT Labs.)

Some compromise in complexity can be reached by using a number of crystals instead of deriving all frequencies from a single crystal. A synthesizer of this type is shown in Fig. 2-23. This is the transmitter section of the AN/ARC-27 and generates frequencies at 0.1-Mc intervals between 225 and 400 Mc. Twenty-two crystals are used (31).

The 10-Mc harmonic generator produces an output every 10 Mc from 200 to 370 Mc. This is done not by direct multiplication but by mixing one group of 10-Mc harmonics with multiples of the ninth harmonic of the 10-Mc oscillator. For example, 310 Mc is generated by taking the third harmonic of 90 Mc and adding this to 40 Mc.

Ten crystals in the 9.15- to 8.25-Mc range provide the 0.1-Mc steps, and these are mixed with the output of the 3.45-Mc oscillator to obtain ten frequencies in the range of 5.7 to 4.8 Mc. Ten more crystals in the 25.7- to 34.7-Mc range provide the 1-Mc steps, and these are mixed with the output of the 5.7- to 4.8-Mc i-f amplifier to produce an output which is mixed with the 200- to 370-Mc spectrum. The particular frequencies used to obtain an output of 341.1 Mc are shown in the boxes in Fig. 2-23.

A synthesizer using a disciplined oscillator is shown in Fig. 2-24. The variable-frequency oscillator operates at the actual desired output frequency, which is anywhere between 2 and 30 Mc in 0.5-kc steps. If any part of the synthesizer fails, operation of the variable-frequency oscillator can continue in the undisciplined condition. The variable-frequency oscillator normally operates as a phase-locked oscillator, and its output is heterodyned down with crystal oscillators to obtain a frequency in the 2- to 3-Mc range. A reference frequency is generated which may be on any 0.5-kc step in the 2- to 3-Mc range by mixing the output of an interpolation oscillator with a frequency generated by one of ten crystals in a crystal oscillator. The various frequencies required to obtain an output of 27,436.5 kc are shown in the boxes in Fig. 2-24. In operation the variable-frequency oscillator is adjusted until it falls within the pull-in range of the phase-locking circuit.

For frequency-shift keying service, the interpolation oscillator is shifted ± 425 cps by the keying voltage. The variable-frequency oscillator will shift by the same amount since it is phase-locked to the output of a mixer which is driven by the interpolation oscillator.

A number of more complex synthesizers have been described in Ref. 31 and elsewhere (38,39).

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R-F POWER AMPLIFIERS

3-1. General Amplifier Considerations. Today's complex communications transmitters utilize many types and varieties of amplifiers. The choice of the type of amplifier depends on many factors. The type of service, the character of the r-f signal to be handled, the operating frequency and range are but a few of the items considered in the choice and design of an amplifier. Actually, one of the prime considerations is not electrical at all but economical. An amplifier is essentially a power converter. A d-c input is converted into an r-f output. This conversion is accomplished at different levels of efficiency for different kinds of amplifiers. In high-power amplifiers the difference in efficiency between one type of amplifier and another might account to a considerable difference in operating costs over the years. Thus the choice of the type of amplifier used in a particular application is an important one, particularly as the power level increases. However, it is nevertheless equally important to use an amplifier of proper electrical characteristics.

Even though the physical appearance of an a-m broadcast power amplifier may be completely different from one of the corresponding power level in a TV transmitter, the amplifier may be electrically the same. Conversely, the characteristics of any given amplifier may be completely altered by a small change in the operating voltages of the tube. To appreciate these differences and to recognize the similarities requires an intimate knowledge of amplifier operation.

3-2. Amplifier Classifications. From the foregoing, the necessity for classifying the many types of amplifier can be readily appreciated. There are many methods of classifying an amplifier. Many of these are self-explanatory. For example, describing an amplifier as a vhf, uhf, or h-f amplifier describes the general frequency range of the amplifier. Similarly, classification of amplifiers according to the power level, type of tube (triode, tetrode), type of service (a-m, f-m, c-w) all provide information and aid in classifying and describing the amplifier. However two widely used methods of classifying amplifiers, which convey more information concerning amplifier characteristics than most descriptive adjectives commonly used, are (1) according to the dynamic operating condi-

tions of the tube and (2) according to the tube connection in the circuit. The former classification is according to the letters A, B, and C. The latter classification leads to terms such as the *grounded-cathode amplifier*, the *grounded-grid amplifier*, and the *grounded-plate amplifier*. The last one is also more commonly known as the *cathode follower*.

Each class of operation and each circuit configuration has its advantages—and disadvantages. No one amplifier is superior in too many respects to any of the others. As a result, all classes of operation are commonly used and intermixed with the various tube connections, in today's complex communications equipment. From the basic definitions that follow these characteristics are not apparent, but they will be made clear in succeeding pages.

3-3. Dynamic Operating Conditions. Class A Amplifiers. Any amplifier that operates such that the plate current flows continuously throughout the electrical cycle is considered as operating under class A conditions. If the grid current flows at any time during the electrical cycle, the subscript 2 may be added to indicate this fact. The subscript 1 indicates that no grid current is flowing. Thus a class A amplifier may be noted as either A_1 or A_2 .

Class AB Amplifiers. If the plate current in an amplifier flows for a period appreciably more than half but less than the entire 360° of the electrical cycle, the amplifier is considered to be operating in class AB. Once again, if grid current flows at any time, the subscript 2 may be added to indicate this fact. If no grid current flows, then a subscript 1 can be added to denote this fact.

Class B Amplifiers. In this type of amplifier, the grid bias voltage is adjusted such that the plate current flows for essentially one half of the electrical cycle, i.e., 180° . Typically the class B amplifier draws grid current; subscripts are not normally used to denote this fact.

Class C Amplifiers. The class C amplifiers are characterized by a plate current that flows in pulses, and which, by definition, must be less than one-half of the electrical cycle. The typical class C amplifier also draws grid current, and subscripts are not normally used to indicate this fact.

3-4. Amplifier Configurations. There are three basic connections for any amplifier. These are based on which element within the tube is made common to both the input and output circuits. The basic configuration for each of these types of amplifiers, i.e., the grounded-cathode, the grounded-grid, and the grounded-plate, is shown in Fig. 3-1. The term *grounded* (which is derived more from common use than from its absolute technical correctness) refers to the a-c impedance of the element to ground or the common point between input and output. The element need not be grounded, but need only be common to both the input and

output. However, in practice, the common lead is usually grounded to the chassis.

3-5. Amplifier Efficiency. One of the important characteristics of a vacuum tube in transmitter applications is the efficiency of the conversion of d-c to r-f power. The d-c power which is not converted to useful r-f power is for the most part converted to heat. This heat represents d-c power that remains unconverted and hence wasted. Thus the

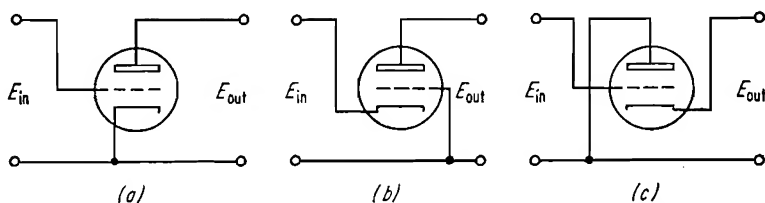


FIG. 3-1. Basic amplifier configurations. (a) Grounded cathode; (b) grounded grid; (c) grounded plate or cathode follower.

result of low efficiency is increased operating cost for power. However, it does not end with this fact alone. Low efficiency compounds itself. The wasted power must be dissipated, which means increased rating for the blowers and/or heat exchangers needed to cool the equipment. The increased dissipation requires tubes with increased power ratings. This also requires increased ratings on all power supply components. Thus, for a given application, and in all but the very lowest power applications, the efficiency must be carefully considered, consistent with the other requirements of the system.

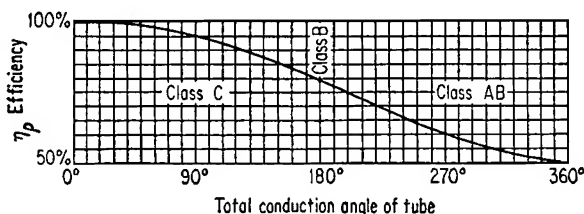


FIG. 3-2. Plate efficiency vs. conduction angle for an amplifier with tuned load. (From Thomas L. Martin, Jr., "Electronic Circuits," p. 452, Prentice-Hall, Inc., Englewood Cliffs, N.J., 1955.)

3-6. Efficiency and Plate-current Conduction Angle. The theoretical efficiency of any amplifier depends on the conduction angle of the plate current. This is illustrated graphically in Fig. 3-2. This curve plots the conduction angle vs. the efficiency of amplifiers operating with a tuned load. This curve is an actual plot of the following expression (28):

$$\text{Efficiency } \% = \frac{\theta_c - \sin \theta_c \cos \theta_c}{\sin \theta_c - \theta_c \cos \theta_c} 50 \quad (3-1)$$

where θ_c is one-half the total conduction angle in radians. From this curve, we can readily see that a tuned class A amplifier has a maximum theoretical efficiency of 50 per cent, the class B has a maximum theoretical efficiency of 78.5 per cent, and the class C can operate at a theoretical efficiency approaching 100 per cent. Practical operating efficiencies are considerably lower than the theoretical maximums listed above. For example, the class A amplifier usually operates with plate efficiencies of

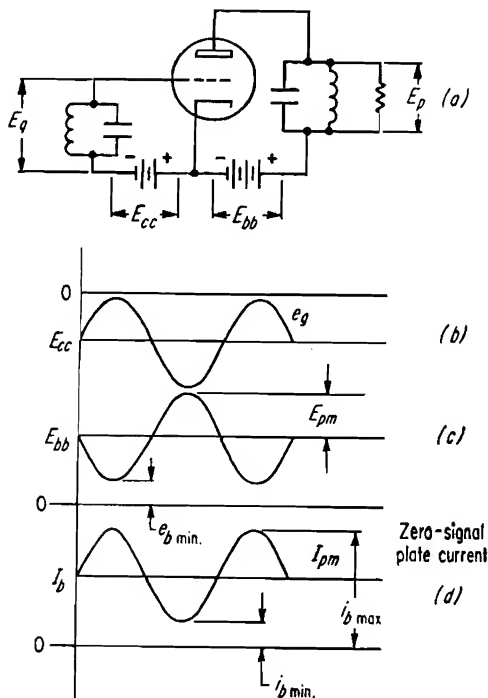


FIG. 3-3. Waveforms in a typical class A tuned power amplifier. (a) Basic amplifier circuit; (b) grid-voltage waveform; (c) plate-voltage waveform; (d) plate-current waveform.

20 to 30 per cent, the class B tuned amplifier usually operates at efficiencies of 60 to 70 per cent, and the class C amplifier operates with efficiencies of 65 to 85 per cent.

3-7. The Characteristics of the Class A Amplifier. Although the class A amplifier exhibits a lower efficiency than other types of amplifiers, it also exhibits some desirable properties which makes its use practical in many applications. The class A amplifier is capable of low amplitude distortion and excellent linearity. These properties makes its use very desirable for applications requiring linear amplification such as in single-sideband (ssb) transmitters. However, the class A amplifier must be

properly designed and operated to achieve this result. It is not inherent nor automatic in this type of amplifier. In general, if the amplifier is not driven to excessive grid current, if the tube is operated over a small portion of its dynamic characteristic, and if the extremes of the characteristic curve are avoided, the class A amplifier will be quite linear. The subject of amplifier linearity is discussed further in Sec. 3-16.

Class A Power Relationships. In Fig. 3-3 the instantaneous conditions for the plate voltage, plate current, and the grid voltage are illustrated for a typical class A amplifier. Assuming the amplifier is operating without distortion, the relationships existing in this amplifier are as follows:

$$E_{pm} = E_{bb} - e_{b\min} = E_{bb} \left(1 - \frac{e_{b\min}}{E_{bb}} \right) \quad (3-2)$$

$$I_{pm} = I_b - i_{b\min} = I_b \left(1 - \frac{i_{b\min}}{I_b} \right) \quad (3-3)$$

$$\text{Output power} = \frac{E_{pm} I_{pm}}{2} = \frac{E_{bb} I_b}{2} \left(1 - \frac{e_{b\min}}{E_{bb}} \right) \left(1 - \frac{i_{b\min}}{I_b} \right) \quad (3-4)$$

$$\text{Plate efficiency} = \frac{\text{power in load}}{E_{bb} I_b} = 0.5 \left(1 - \frac{e_{b\min}}{E_{bb}} \right) \left(1 - \frac{i_{b\min}}{I_b} \right) \quad (3-5)$$

$$\text{Load resistance} = R_L = \frac{E_{pm}}{I_{pm}} = \frac{E_{bb}}{I_b} \left[\frac{1 - \frac{e_{b\min}}{E_{bb}}}{1 - \frac{i_{b\min}}{I_b}} \right] \quad (3-6)$$

Class A Amplifier Efficiency. Although the efficiency of any amplifier depends on the conduction angle, the efficiency of any amplifier with a given conduction angle can vary from the maximum specified in the curve of Fig. 3-2 all the way to 0 per cent efficiency. Thus the efficiency at any given angle depends on the operating conditions of the tube and the magnitude of the driving signal. From Eq. (3-5), we can see that, if the operating conditions are chosen to make the ratios $e_{b\min}/E_{bb}$ and $i_{b\min}/I_b$ zero, the efficiency of the amplifier would be 50 per cent, the maximum theoretically possible. This expression also indicates that, for a practical approach to this theoretical efficiency, it is necessary to keep both $e_{b\min}/E_{bb}$ and $i_{b\min}/I_b$ small. In practice, the ratio of $i_{b\min}/I_b$ can be kept quite small. As a matter of fact, it can be made to approach zero if some distortion can be tolerated. The problem lies in attempting to reduce the ratio of $e_{b\min}/E_b$ to very low levels. It is difficult to operate a tube so that it will draw a high value of $i_{b\max}$ at extremely low values of $e_{b\min}$. In practice, a compromise must be reached, which will depend on the distortion and nonlinearity that can be tolerated and the efficiency desired.

Class A Amplifier Operating Conditions. Usually the operating conditions for a tube are given in the tube manuals or manufacturer's literature. However, when these are not available, or it is desired to operate the tube under different conditions, the operating conditions for the tube must be estimated. From the curves of the tube, we determine the value of $e_{b\min}$ that will cause a plate current to flow that is approximately twice the rated plate current at zero grid voltage. This applies if no grid current is to flow, i.e., the amplifier is to operate as class A_1 . If the amplifier is to operate as a class A_2 amplifier with grid current, then the value of $e_{b\min}$ is determined as above for the appropriate value of positive grid driving voltage. The load resistance can then be calculated from Eq. (3-6). The results can be refined by drawing the load line on the characteristic curve for the load resistance just calculated.

If the tube is a pentode or a tetrode, the screen grid is normally chosen to result in a plate current of twice the rated value with a grid-bias voltage of zero. The grid-bias voltage is then set to a value that will result in the rated plate-current flow.

Maximum power will be obtained from the amplifier if the load resistance is made equal to the plate resistance of the tube. For load resistances either higher or lower, the power output will decrease as shown in Fig. 3-4. However, if minimum distortion is a prime consideration, then operation with such a value of load resistance may not be desirable. The distortion level decreases, and the linearity increases when the load

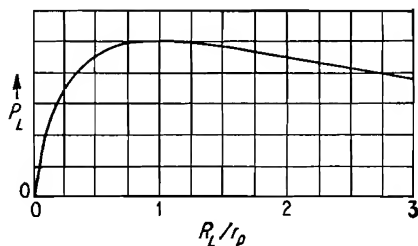


FIG. 3-4. Effect of load resistance on power output. (From Thomas L. Martin, Jr., "Electronic Circuits," p. 408, Prentice-Hall, Inc., Englewood Cliffs, N.J., 1955.)

current is to flow, i.e., the amplifier is to operate as class A_1 . If the amplifier is to operate as a class A_2 amplifier with grid current, then the value of $e_{b\min}$ is determined as above for the appropriate value of positive grid driving voltage. The load resistance can then be calculated from Eq. (3-6). The results can be refined by drawing the load line on the characteristic curve for the load resistance just calculated.

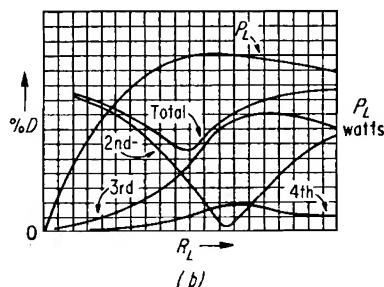
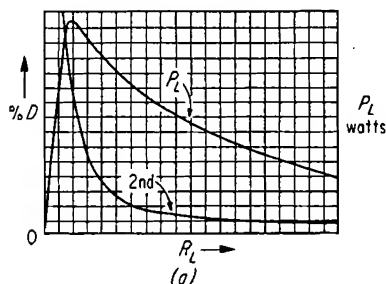


FIG. 3-5. Principal harmonics generated in triodes and pentodes as a function of load resistance. (a) Power output and second-harmonic distortion in a triode as a function of load resistance; (b) power output and harmonic distortion in a pentode as a function of load resistance. (From Thomas L. Martin, Jr., "Electronic Circuits," p. 411, Prentice-Hall, Inc., Englewood Cliffs, N.J., 1955.)

resistance for a triode is increased. Basically, the distortion in a triode is the result of the variable plate-resistance characteristic. By making the load resistance higher, the variations in the plate resistance compared to the total resistance become less—hence less distortion results.

Pentodes and tetrodes require a load lower than the plate resistance for best linearity and minimum distortion. The general distortion-vs.-load-resistance characteristic for pentodes and triodes is illustrated in Fig. 3-5.

Maximum power output for the class A amplifier is obtained by driving the tube with larger signals. The power output is proportional to the square of the driving signal as shown in Fig. 3-6. As indicated previously, this will also result in increased distortion which may not be tolerable, depending on the system requirements.

An aid in selecting a power-amplifier tube is the product of the μ and transconductance (g_m). The higher the product of the tube μ and g_m , the greater will be the power output from the tube. This figure of merit is based on the following expression for power output (Ref. 28, p. 407).

$$P_L = \mu g_m \frac{R_L/r_p}{1 + R_L/r_p} \frac{E_g^2}{2} \quad (3-7)$$

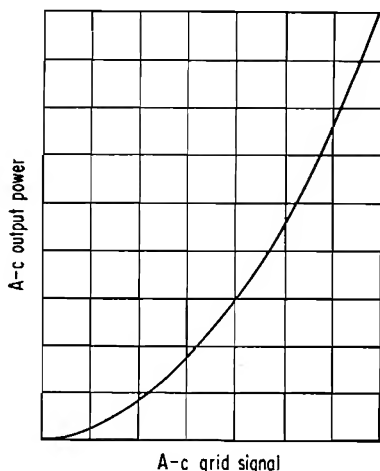


FIG. 3-6. Amplifier power output rises as the square of the a-c grid signal.

3-8. The Characteristics of the Class B Amplifier. The class B amplifier is characterized by a higher efficiency compared to the class A amplifier. This is due to the shorter conduction angle of the class B amplifier. (This was discussed in detail in Sec. 3-6.) The class B amplifier is also capable of being adjusted to produce a linear relationship between the input and output signals. The two advantages of the class B amplifier—i.e., high efficiency and good linearity—usually are the reason for its use in transmitting equipment. This is particularly true at the higher power levels where amplifier efficiency becomes increasingly important.

In practice, the grid bias for a class B amplifier is adjusted for cutoff, i.e., the grid bias voltage is adjusted to a value such that the plate current is zero or nearly so. Thus the plate current will flow only during the time the input signal is going through its positive alternation. As

the input signal goes through the negative half of the alternation, no plate current can flow since the tube is already cut off by the grid bias voltage. Thus the plate current only flows for one-half of every input cycle in the class B amplifier. At first glance, the class B amplifier might appear to be producing a severely distorted output signal, since only half of the input signal results in any plate-current flow. However, if the amplifier has a tuned LC circuit in its output, the flywheel effect due to resonance of the circuit produces the "missing" half of the cycle, even when no plate current is flowing.

In order to obtain the maximum efficiency from the class B amplifier, it is necessary to drive the tube so that the plate-voltage swing is as large as is practicable. The greater the plate swing, i.e., the smaller $e_{b\min}$ is, the higher will be the efficiency of the amplifier. In order to accomplish the required plate-voltage swing, the grid must be driven positive in the majority of class B amplifier designs. The resultant grid current places a varying load on the driver stage. This requires that the r-f regulation of the driver be considered in the over-all design of the transmitter. This fact plus the fact that the driver must deliver the power that causes this current to flow complicates the practical design of a class B amplifier and its application in a transmitter.

If the class B amplifier is used to handle ssb or TV signals, then the plate current will vary with the audio or video level being transmitted. This varying plate current requires that the power supply regulation be carefully considered. In the case of the class B amplifier handling video a-m signals, the power supply must exhibit a low impedance to all the video modulating frequencies to prevent ringing and similar defects in the final picture.

If the amplifier is handling a voice or similar a-m r-f carrier, the average plate current as seen on the plate-current meter will remain constant. Actually, the instantaneous plate current will vary with the a-m r-f signal, but, if the modulation is symmetrical, the current peaks will average out the current valleys. Thus the plate current will appear to be constant on the plate-current meter.

Class B Power Relationships. The instantaneous conditions for the plate voltage, the plate current, and the grid voltage are illustrated for a typical class B amplifier in Fig. 3-7.

The relationships existing in the class B amplifier are as follows:

$$P_{in} = E_{bb}I_b = \frac{2\sqrt{2}}{\pi} E_{bb}I_{pm} \quad (3-8)$$

$$P_{out} = I_{pm}^2 R_L = \frac{i_{b\max}^2 R_L}{8} \quad (3-9)$$

$$\text{Efficiency} = n = 78.5 \frac{E_{pm}}{E_{bb}} \quad \text{per cent} \quad (3-10)$$

Class B Amplifier Operating Conditions. A close estimate of the operating conditions for any class B or class AB₂ amplifier can be calculated from the graphical method outlined in Sec. 3-10. Since this method is also applicable to other classes of amplifiers, one should be aware of the peculiarities of the class B amplifier, and to be more specific, of the amplifier to be designed. For example, if the stage is to handle an

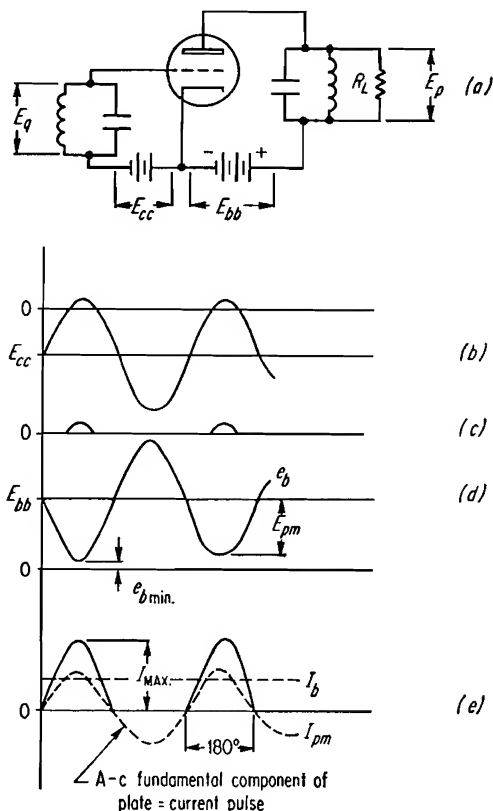


FIG. 3-7. Waveforms in a typical class B tuned power amplifier. (a) Basic amplifier circuit; (b) grid-voltage waveform; (c) grid current; (d) plate-voltage waveform; (e) plate-current waveform.

a-m signal, then the amplifier will have to be capable of delivering the peak power at the peak of the modulated r-f drive signal. The design must be made on the basis of the peak power output conditions rather than the average r-f carrier conditions that may exist. For an amplifier which will handle an ssb signal, then the amplifier linearity may be an even greater problem and will have to be carefully considered when choosing the operating conditions. It may be necessary to allow some idling plate current to flow rather than set the grid bias exactly to plate-

current cutoff. This will help in minimizing the distortion generated by the amplifier. Additional design considerations imposed by the circuit configuration or operating conditions must also be considered. Additional information can be found in Secs. 3-16 and 3-17 on linear amplifiers and wideband amplifiers.

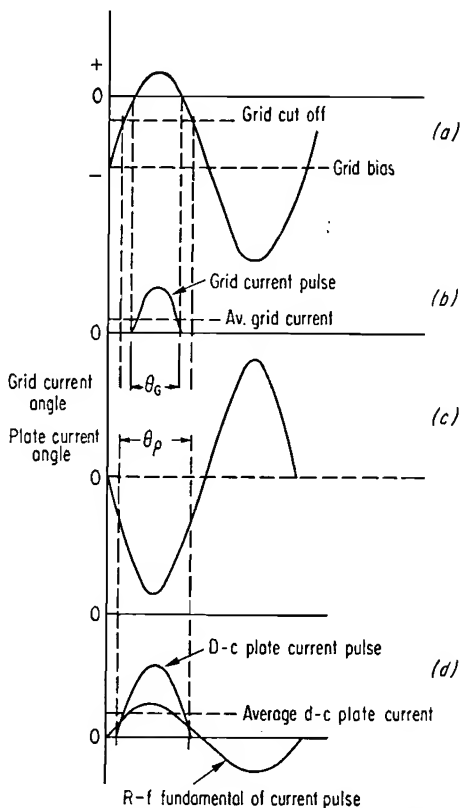


FIG. 3-8. Voltage and current relationships in class C amplifier. (a) Grid voltage; (b) grid current; (c) plate voltage; (d) plate current.

The simplified method of determining the operating conditions for an amplifier described in Sec. 3-10 is based on the constant-current curves of the tube. These are usually available in the manufacturer's literature or on request from the manufacturer. In those cases where this information is not available, the curves which are available can be redrawn to obtain the constant-current curves.

3-9. The Characteristics of the Class C Amplifier. The class C amplifier is characterized by its relatively high efficiency (compared to other classes of amplifier operation). This point was discussed in

Sec. 3-5. This high efficiency is primarily due to its shorter plate-current conduction angle which permits the plate current to flow when the plate voltage is at its lowest value (see Fig. 3-8). The lower the voltage drop across the tube when it is conducting, the higher will be the efficiency of the amplifier. However, to make the plate conduction angle short (less than 180°) requires that the grid-bias voltages be adjusted to beyond plate-current cutoff. This results in the nonlinear transfer characteristic for the class C amplifier. However, the class C amplifier does exhibit a linear change in the r-f voltage output for a change in the d-c plate voltage. This characteristic permits the class C amplifier to be plate-amplitude-modulated.

The plate and grid currents of a class C amplifier are periodic pulses, the duration of which is typically in the order of 120 to 140° . For this reason, the average plate and grid currents, etc., cannot be directly calculated as in other classes of amplifiers. The operating conditions of a class C amplifier can best be approximated by either analytical or graphical means. The graphical method gives sufficient accuracy for most design work and permits the rapid evaluation of several amplifier designs in a very short time. This method is presented in Sec. 3-10.

3-10. A Simplified Method of Amplifier Design. A method of making a close estimate of the performance of an amplifier operating as a class B, C, or as a frequency multiplier is described below. This method requires the use of a simple "tube performance computer" which can either be obtained from the Eitel-McCullough Corporation, San Bruno, California, or can be made from an 8- by 10-in. piece of clear plastic sheet. The details of the tube performance computer are shown in Fig. 3.9.

The basic steps in using the tube computer are outlined in Fig. 3-10. This requires selecting d-c plate and grid voltages, being guided by the typical operating values given on the technical data sheet for the tube type and by general experience. Next a suitable "operating line" must be chosen for the tube type (plotted on the grid-voltage-plate-voltage curves).

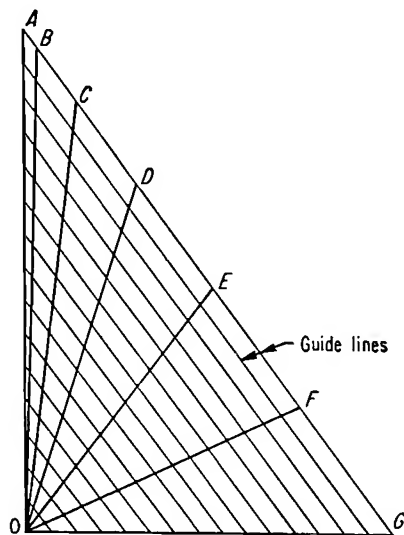


FIG. 3-9. The "tube-performance computer" used in the simplified method of amplifier design. (Courtesy of Eitel-McCullough, Inc.)

1. Mark point of d-c plate voltage and d-c grid bias.
2. Mark point of peak plate current in low plate-voltage region (this is about four times d-c plate current).
3. Draw a straight line between the points selected in steps 1 and 2. This is the operating line.
4. Place the computer on the curve sheet with the guide lines parallel to the operating line." Make *OG* line of computer go through point of step 1. Make *OA* line go through point of step 2.
5. Read the current values where operating line crosses *OA*, *OB*, *OC*, *OD*, *OE*, and *OF*.
6. Put the values in formulas of Fig. 3-11 as *A*, *B*, *C*, *D*, *E*, and *F*. For detailed instructions see Sec. 3-10 of text.

FIG. 3-10. Outline of procedure for use of the tube computer.

"The computer, when properly placed over this operating line, enables one to obtain instantaneous values of the currents flowing at every 15° of the electrical cycle.¹ The formulas given in Fig. 3-11 for use with the computer were derived by Chaffee to give the various average and harmonic components of the resulting currents. Knowing these current component values and the r-f voltage values which are indicated by the use of the computer, one can readily calculate the complete performance of the tube.

D-c (meter reading)	$= \frac{1}{12}(0.5A + B + C + D + E + F)$
Peak fundamental radio frequency	$= \frac{1}{12}(A + 1.93B + 1.73C + 1.41D + E + 0.52F)$
Peak second-harmonic radio frequency (approx.)*	$= \frac{1}{12}(A + 1.73B + C - E - 1.73F)$
Peak third-harmonic radio frequency (approx.)*	$= \frac{1}{12}(A + 1.41B - 1.41D - 2E - 1.41F)$
Output power	$= \frac{1}{2} \text{ peak r-f plate current} \times \text{peak r-f plate voltage}$
Driving power	$= \text{d-c grid current} \times \text{peak r-f grid voltage}$

* Use only for tetrodes or pentodes; approximate only.

FIG. 3-11. Equations for use with the tube performance computer shown in Fig. 3-9 and the constant-current tube curves to obtain the plate, grid, and screen current as well as the output and driver power for the amplifier.

"The fundamental methods of making such computations and the considerations necessary to stay within the ratings of the tube types and accomplish various forms of modulation have been covered in the literature. The method for the case of harmonic amplifier service is approximate and should only be used for tetrode and pentode tubes, where the plate voltage has little effect on the amount of plate current flowing. A more exact method, showing that for harmonic operation the operating line is a simple Lissajous figure, has been described by Brown.

¹ The remainder of this section is quoted from Ref. 19.

"The results of using this computer for power-amplifier service can be applied in combination with other methods described to give good accuracy with simpler procedures. The resulting accuracy is well within the normal variations of tube characteristics due to the normal variation in manufacturing dimensions of a tube. Since the published tube curves are only typical of the characteristics to be expected from a particular tube type, the calculated performance is well within the values expected when different tubes of a given tube type are operated under the assumed conditions.

"Example Showing Detailed Use of the Eimac Tube Performance Computer. Let us say we have an Eimac 4-65A tetrode and want to make it work effectively. Also let us say we have a 2,000-volt d-c plate supply available. It is also desired to use this tube as a class C amplifier.

"Within frequency limits, we know a tube should be able to run in class C amplifier service with about 75 per cent efficiency, or in other words, to convert 75 per cent of the d-c plate input power into r-f output power. The difference, or 25 per cent of the input power, is dissipated or lost as heat on the plate of the tube. The d-c plate input power is then about four times the power dissipated on the plate.

"The 4-65A tetrode has a maximum rated plate dissipation of 65 watts. So, to illustrate performance near the maximum rating, we'll choose an input power four times the plate dissipation, or 260 watts per tube. At 2,000 volts the plate current per tube must be 130 ma. It is usual practice in the case of tetrodes and the medium- or low- μ mu triodes in class C amplifier service for the d-c grid-bias voltage to be roughly two or three times the grid voltage necessary to cut off the flow of plate current. By referring to the curves of the 4-65A in Fig. 3-12, we decide to use a d-c grid-bias voltage of -120 volts.

"Let us now locate the operating line on the constant-current curves of the 4-65A. First mark the point where the d-c grid-bias and d-c plate voltage cross. The operating line must go through this point. Call it point 1. Next we must decide what the peak value of the plate current of the tube must be and how low we can let the instantaneous value of the plate voltage go when the tube is passing this much current. This is necessary in order to locate the other end of the operating line, point 2.

"The peak value of plate current usually runs about four times the d-c plate current. The minimum value of instantaneous plate voltage is usually set by the fact that if the voltage is too low, the grid and screen currents will be needlessly high, and also little will be gained as far as output power is concerned. The minimum value of plate voltage is usually in the region where the plate constant-current curves bend upward (see Fig. 3-12). (In the case of the triode this is near the 'diode line' or the line where the instantaneous grid and plate voltages are equal.)

The practical procedure in calculating tube performance is to arbitrarily choose point 2 and complete the calculations. Then try other locations of point 2, complete the calculations, and compare the results.

"In the case of the 4-65A, let us choose a peak value of plate current about four times the d-c plate current of 130 ma, or 500 ma. Let us choose a minimum instantaneous plate voltage of 250 volts and thus fix the upper end of the operating line. Next locate this point on the tube curves; this is point 2 on Fig. 3-12. (The plate currents which flow at various combinations of plate and grid voltages are shown by the plate-current lines. The value of current for each line is noted. In between

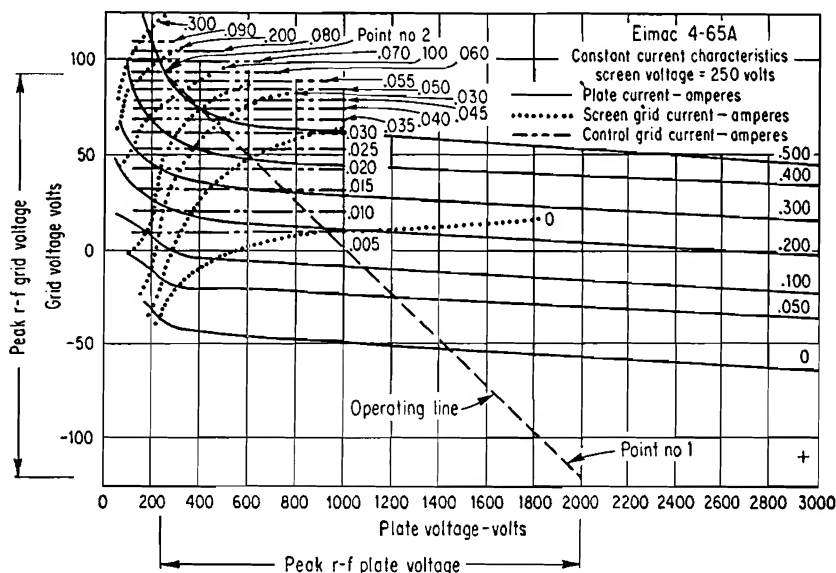


FIG. 3-12. Characteristic curve for 4-65A tube used in example described in Sec. 3-10. (Courtesy of Eitel-McCullough, Inc.)

values can be estimated closely enough for our purposes.) Now draw a straight line between points 1 and 2. This line is the operating line and shows the current and voltage values for each part of the r-f cycle when current is being taken from the tube. (The nonconducting half of the r-f cycle would be shown by extending this line an equal distance on the opposite side of point 1. However, there is little use in doing so because no current flows during this half of the cycle.)

"The Eimac tube-performance computer can now be used to obtain the meter readings and power values from this operating line. Place the computer on the constant-current curve sheet so that the guide line of the computer are parallel with the operating line. Now slide the computer about without turning it until the line *OG* passes through the

d-c voltage point 1 and the line OA passes through the peak current point 2. Make sure the guide lines are still parallel to the operating line.

"Note that the lines OB , OC , OD , OE , and OF of the computer all cross over the operating line.

"At each point where the lines OA , OB , etc., cross the operating line we need to determine the instantaneous values of plate current and grid current (and screen current if a tetrode or pentode is used) which is flowing at that particular moment in the r-f cycle. Later, from these key values of current, we will calculate the values of d-c plate current and grid current (and screen current) as well as the r-f components of the plate current.

"At each of these points, where the instantaneous current values are to be determined, a mark should be made on the constant-current curve sheet of the tube. By noting where this mark lies with respect to the plate-current curves, one can estimate the value of plate current flowing at this part of the cycle. Next, the location of this mark with respect to the control grid curves is noted and a value of grid current is estimated. Finally, by referring the mark to the screen-grid curves, if the tube is a tetrode or pentode, a value of screen current is noted. These current values should be listed for each point where the lines OA , OB , etc., cross the operating line so that they can be combined later to calculate the various tube currents. At points where OF and OE cross, the current values are often zero.

"Now in the example chosen, let us read off the instantaneous plate-current values where these cross the operating line. At the point where the line OA crosses the operating line the plate current is 500 ma. Where OB crosses the operating line, the plate current can be estimated as 510 ma since the point is about one-tenth of the way from the 500-ma line to the 600-ma line. At OC the plate current is 460 ma; OD , 290 ma; OE , 75 ma; OF and OG , 0 ma. Similarly we can estimate the instan-

TABLE 3-1. VALUES OF LINES IN FIG. 3-11

Crossing of line	Simplified name in formulas	Instantaneous values of currents, ma		
		Plate	Screen	Control grid
OA	A	500	165	60
OB	B	510	100	50
OC	C	460	25	30
OD	D	290	5	14
OE	E	80	0	0
OF	F	0	0	0

taneous screen current at the crossing of OA and the operating line as 165 ma, and the instantaneous grid current at 60 ma. Values are read for the other crossings and written down. These values are put in simple columns as shown in Table 3-1.

"Now in order to obtain the d-c value of plate, screen, and control grid currents, the formula (see Fig. 3-11) says to add up the above values but use only one-half the A values (giving 250 ma for plate, 82 ma for screen, and 30 ma for grid), and then divide by 12 as is done in Table 3-2. Thus d-c plate-meter current is 132 ma, screen current is 18 ma, and control grid current is 10 ma.

TABLE 3-2. D-C METER READING = $0.5A + B + C + D + E + F$

Plate, ma	Screen, ma	Control grid, ma
250	82	30
510	100	50
460	25	30
290	5	14
80	0	0
Total 1,590 ma	212 ma	124 ma

"Now, to calculate the r-f output power, it is necessary to use the formula for the peak r-f current which is present in the tube plate current. Since we are using the tube as a straight r-f amplifier, we use the formula for "peak fundamental radio frequency" as shown in Fig. 3-1. If we were estimating the performance of a doubler or a tripler, we would use the formula for "peak second harmonic radio frequency" or "peak third harmonic radio frequency."

"From the computer we see that the formula for the peak fundamental r-f current is:

$$\frac{1}{12} (A + 1.93B + 1.73C + 1.41D + E + 0.52F)$$

$$\begin{array}{rcl} A & = & 500 \text{ ma} \\ 1.93B & = & 1.94 \times 510 = 985 \text{ ma} \\ 1.73C & = & 1.73 \times 460 = 796 \text{ ma} \\ 1.41D & = & 1.41 \times 290 = 409 \text{ ma} \\ E & = & 80 \text{ ma} \\ 0.52F & = & 0 \end{array}$$

Total 2,770 ma

$$\text{Peak fundamental current} = \frac{1}{12} (\text{total}) = \frac{2,770}{12} = 230 \text{ ma}$$

"We now have the various current values. In order to calculate the powers involved, it is necessary to know not only the d-c voltage values but the greatest amount each voltage swings away from the d-c value.

This is known as the *peak value* of the r-f voltage. Because the plate voltage swings from 2,000 volts down to 250 volts, the peak r-f voltage is the difference, or 1,750 volts. Similarly the grid voltage must rise and fall between the operating points 1 and 2, or from -125 volts to +95 volts. This is a peak swing of 220 volts, and the peak r-f grid voltage is 220 volts.

"Let us now use the formulas for output power and driving power:

$$\text{Output power} = \frac{1}{2} \text{ peak r-f plate current} \times \text{peak r-f plate voltage}$$

We found the peak r-f plate current to be 230 ma or 0.230 amp, and the peak r-f plate voltage to be 1,750 volts. Therefore,

$$\text{Output power} = \frac{1}{2} \times 0.230 \times 1,750 = 201 \text{ watts}$$

$$\text{Input power} = \text{d-c plate current} \times \text{d-c plate voltage}$$

$$= 0.132 \times 2,000 = 264 \text{ watts}$$

$$\text{Plate dissipation} = \text{d-c input power} - \text{r-f output power}$$

$$= 264 - 201 = 63 \text{ watts}$$

$$\text{Efficiency} = \frac{\text{r-f power}}{\text{d-c input power}}$$

$$= \frac{201}{264} = 76\%$$

$$\text{Driving power} = \text{d-c grid current} \times \text{peak r-f grid voltage}$$

$$= 0.010 \times 220 = 2.2 \text{ watts}$$

"The power consumed by the bias source is simply the product of the d-c grid current and the d-c grid voltage, or $0.010 \times 120 = 1.2$ watts. The difference between the driving power and the power consumed by the bias source is the power dissipated on the control grid, or

$$2.2 - 1.2 = 1.0 \text{ watts}$$

"The power dissipated on the screen grid is simply the product of the d-c screen current and the d-c screen voltage, because the screen grid has no impedance between it and the d-c screen supply. Thus it is $0.018 \times 250 = 4.5$ watts. The performance of the tube can now be summarized as seen in Table 3-3."

TABLE 3-3. TUBE PERFORMANCE SUMMARY*

D-c plate voltage.....	2,000 volts	Driving power.....	2.2 watts
D-c screen voltage.....	250 volts	Grid dissipation.....	1.0 watts
D-c grid voltage.....	-250 volts	Screen dissipation.....	4.5 watts
D-c plate current.....	132 ma	Plate power input.....	264 watts
D-c screen current.....	18 ma	Plate power output.....	201 watts
D-c grid current.....	10 ma	Plate dissipation.....	63 watts
Peak r-f voltage.....	220 volts		

* Quoted from Ref. 19,

3-11. The Grounded-cathode Amplifier. The grounded-cathode amplifier configuration is used almost exclusively for all the r-f power amplifiers in the low-, medium-, and high-frequency ranges. Its use extends well into the vhf region. However, at these frequencies, the advantages of the grounded-grid amplifier becomes increasingly apparent. In general, the grounded-cathode amplifier is characterized by both a higher input impedance and a higher output capacitance than the grounded-grid amplifier.

In the grounded-cathode amplifier, the signal is placed between the grid and cathode, and the output is taken between the plate and cathode as shown previously in the basic circuit of Fig. 3-1. Thus there is no intentional coupling between the input and output circuits of the amplifier. As a result the grounded-cathode amplifier is capable of a high degree of isolation between the input and output.

The drive power requirements of the grounded-cathode amplifier used in ordinary amplifier service is lower than the equivalent power-rated grounded-grid amplifier. The drive for the grounded-cathode amplifier is necessary only to provide for the proper operation and excitation of the tube. None of the driver power is passed along to the output. Thus there is no interaction between the input and the output in the grounded-cathode amplifier. If any interaction does exist, it indicates improper or lack of neutralization. If no grid current flows, as is the case if the tube is operated according to class A_1 or AB_1 , then the only drive power necessary is that required to overcome the losses within the tube structure and in the tuned circuit elements in the input circuits to the tube. Calculation of the operating parameters for the grounded-cathode amplifier operating as a class A, B, or C will be found in the appropriate section according to the class of the amplifier.

3-12. The Grounded-grid Amplifier, General Characteristics. The grounded-grid amplifier finds its greatest use in the vhf and uhf region, for it is at these frequencies that the advantage of this circuit configuration become apparent.

The basic configuration of the grounded-grid amplifier was shown in Fig. 3-1. In this circuit, the grid need not necessarily be placed at d-c ground potential. The term *grounded grid* refers to the fact that the impedance between the grid and ground is negligible at the frequencies involved in the amplifier operation. Thus, a d-c potential may exist between the grid and ground for biasing purposes.

Since the grid is essentially at a-c ground, it is apparent that it will also act as a shield, functioning in much the same manner as the screen grid in a tetrode. This characteristic of the grounded-grid amplifier permits a given triode to be operated at a much higher frequency without neutralization than if it were used in a grounded-cathode amplifier stage.

In the grounded-grid amplifier, the feedback capacitance within the tube between the input and output is simply the capacity that exists between the cathode and the plate. In the grounded-cathode amplifier, this feedback capacitance is the capacity that exists between the control grid and the plate. This latter capacitance is usually 25 times the feedback capacitance existing in the grounded-grid amplifier. For this reason, the frequency range of the tube may be extended considerably if it is used as a grounded-grid amplifier. This assumes that care is taken externally in the placement and isolation of the input and output circuits to prevent feedback through this path.

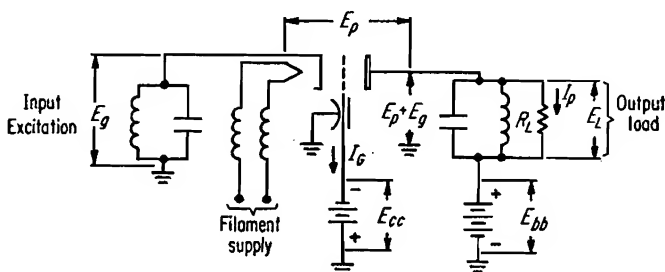


FIG. 3-13. Basic circuit of grounded-grid tuned power amplifier.

A second characteristic apparent from the basic grounded-grid amplifier circuit of Fig. 3-13 is that the driver excitation acts in series with the r-f voltage developed between the plate and cathode of the amplifier tube. These two voltages add, since they are in phase with each other, and act together to supply the output voltage across the load. Thus, if the rms value of the r-f voltages provided by the drive is E_g and that developed by the tube is E_p , then the voltage across the load is $E_g + E_p$.

"If I_p is the fundamental component of the plate current 180° out of phase with E_p and if I_g is the fundamental component of grid current in phase with E_g , the following relations then hold (Ref. 6, p. 138):

$$\text{Power delivered to load circuit} = (E_p + E_g)I_p \quad (3-11)$$

$$\text{Power delivered by output tube} = E_p I_p \quad (3-12)$$

$$\text{Power transferred to load circuit} = E_g I_p \quad (3-13)$$

$$\text{Power delivered by driver circuit} = E_g (I_p + I_g) \quad (3-14)$$

$$\text{Power absorbed by output tube} = E_g I_g \quad (3-15)''$$

The above relationships bring out another important characteristic of the grounded-grid amplifier, that is, the driving power required for grounded-grid operation is considerably greater than the equivalent tube operated in a cathode-grounded circuit. As a matter of practical fact, the increased drive power required is usually three to ten times greater than the same tube operated as a cathode-grounded amplifier. This

may appear as a severe disadvantage on first consideration; however, the relationships stated above reveal that most of the driving power is passed on to the output. The actual power absorbed in the control grid is the same, regardless of the circuit configuration. Thus a typical grounded-grid amplifier and its driver will be no less efficient than the equivalent amplifier chain with a grounded-cathode output stage, if efficiency is considered from the over-all viewpoint. The two amplifier chains will remain on a par if they are operated at frequencies low enough so that the output tube capacity of the amplifier is not the only capacity in the tank circuit of the output. This is usually true at high frequencies and lower.

The grounded-grid amplifier is further characterized by its lower output capacity as compared to the grounded-cathode amplifier. The output capacity is approximately one-half that of the typical capacity neutralized grounded-cathode amplifier. The output capacity of an amplifier operating in the vhf region and up is usually the limiting factor which determines the power output available from the tube and hence, indirectly, the conversion efficiency of the amplifier. This is particularly true in wideband applications such as TV. For example, in a single tuned output stage for an amplifier, the relationship between the load resistance, bandwidth, and output capacity is

$$BW = \frac{1}{2\pi R_L C} \quad (3-16)$$

Thus for a given bandwidth requirement, and with a fixed output capacity, the only variable left is the load resistance. The higher the output capacity, the lower will be the required load resistance to fulfill a given bandwidth requirement. For a given plate-current flow in a tube, the higher the load resistance, the higher will be the power output in the load resistance. Since the grounded-grid amplifier does exhibit a lower output capacity, it can be safely concluded that it will also deliver more output power for a given bandwidth than will its grounded-cathode counterpart.

In narrow-band work, at very high frequencies, the lower output capacity has the advantage that the circulating kva in the output tank circuit is reduced by the same factor as the output capacity. Thus for the same mechanical construction in the output circuit, there will be fewer losses, and hence higher over-all efficiency for the grounded-grid configuration.

*Grounded-grid-amplifier Design Considerations.*¹ "Let it be required to design a grounded-grid stage utilizing a 9C21 triode. The tube will

¹ Quoted from Ref. 6, p. 140.

be used in the circuit shown in Fig. 3-13. The manufacturer's data show the following typical operating conditions for normal cathode-grounded circuits.

D-c plate voltage.....	17,000 volts
D-c grid voltage.....	-1,600 volts
Peak r-f grid voltage.....	2,200 volts
D-c plate current.....	7.9 amp
D-c grid current.....	0.9 amp
Driving power.....	1,800 watts
Power output.....	100 kw

"It is first necessary to obtain the plate-voltage swing E_p . This can be estimated from the fact that, in a properly excited class C amplifier, the plate voltage will swing down to the value of the peak positive grid voltage. Because the value of the latter is 2,200 volts minus 1,600 volts or 600 volts,

$$E_p \text{ is } 0.707 (17,000 - 600) = 11,600 \text{ volts rms}$$

"Next the fundamental components of the plate current and of the grid current must be obtained, using the power-output and driving-power values given above.

$$I_p = \frac{100,000}{11,600} = 8.63 \text{ amp}$$

$$\text{and } I_g = \frac{1,800 \sqrt{2}}{2,200} = 1.16 \text{ amp}$$

The output power and the driving power of the grounded-grid stage can now be calculated from the relations given earlier. We obtain

$$\begin{aligned} \text{Power output} &= (E_p + E_g)I_p \\ &= \left(11,600 + \frac{2,200}{\sqrt{2}} \right) 8.63 = 113 \text{ kw} \end{aligned}$$

$$\begin{aligned} \text{Driving power} &= E_g(I_p + I_g) \\ &= \frac{2,200}{\sqrt{2}} (8.63 + 1.16) = 15.2 \text{ kw} \end{aligned}$$

"It is possible to change the power output to some extent by varying the grid bias and the grid swing. For example, if the bias should be changed from -1,600 volts to -1,200 volts and the grid swing by a like amount, i.e., to 2,600 volts, the power output would be 116 kw."

Modulation Characteristics of the Grounded-grid Amplifier. In most transmitter applications, the grounded-grid amplifier is also used as a linear amplifier, and hence is unmodulated. However, if it becomes necessary to plate-modulate an amplifier using a grounded-grid stage, it will also be necessary to modulate the source of excitation. If the

preceding stage is a grounded-cathode stage, this stage and the grounded-grid-stage should be plate-modulated simultaneously. However if the driver consists of a series of cascaded grounded-grid stages, it will be necessary to modulate two or more of the exciter stages as well as the grounded-grid stage under consideration. The amount of stages that should be modulated depends on the distortion that can be tolerated and the maximum percentage of modulation. This is necessary because

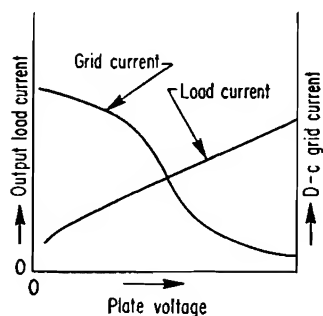


FIG. 3-14. Plate-modulation characteristics of grounded-grid amplifier. (From A. A. Spitzer, *Grounded Grid Power Amplifiers*, *Electronics*, April, 1946.)

the r-f driving voltage is in series with the amplifier in the grounded-grid amplifier.

If the stage is modulated alone, a characteristic curve similar to that shown in Fig. 3-14 would be obtained. From this curve we can see that, if the plate voltage were dropped to zero during the modulation cycle, some amount of output r-f power would still appear due to the driver. If the plate voltage were made negative by the amount of the peak of the r-f driving voltage, then the r-f output would drop to zero. However, this would still result in the nonlinear modulation characteristic illustrated in the curve of Fig. 3-14 at the

lower plate voltages. Thus, for 100 per cent modulation of the grounded grid, the usual method is to modulate the source of excitation as well.

Filament-supply Isolation in the Grounded-grid Amplifier. A problem peculiar to the grounded-grid amplifier is that of heating the filament. Since the r-f drive power is introduced between the filament (or cathode) and ground, the filament is above ground at an r-f voltage. The problem is then how to supply the filament power to the tube without loading down the r-f input circuit. For low-power transmitting tubes, with a low filament current, isolating the filament from its supply through a pair of r-f chokes is no problem. However the larger tubes draw currents in the hundreds of amperes. The r-f choke in these situations becomes a bulky item.

In amplifiers operating up to approximately 25 Mc, the filament can be supplied through a low-capacity filament transformer. The capacity to ground added by the transformer can be compensated for in the r-f input network.

For amplifiers operating in the vhf region, the filament can be supplied through a quarter-wavelength section of the transmission line. This transmission-line section can also form part of the input matching network. In many cases, the frequency of operation is high enough and the tube capacitance great enough so that the greater portion of a quarter

wavelength is within the tube structure. In these cases, the isolation can be made through any odd multiples of a quarter wavelength. This and other systems of filament isolation for the grounded-grid amplifier is shown in Fig. 3-15.

Tetrodes in Grounded-grid Amplifiers. While the discussion which has preceded has been centered on triodes, tetrodes can be used quite satisfactorily in grounded-grid circuits. The grounded-grid configuration requires that the screen-grid bypassing capacitor be returned to ground and not to the cathode. If the screen-grid capacitor was returned to the cathode, the shielding between the input and output circuits would be

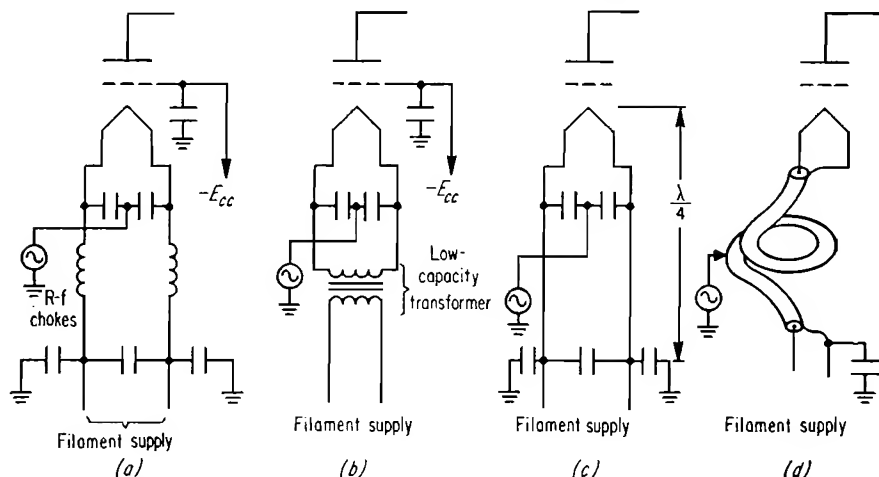


FIG. 3-15. Methods of supplying filament power for the grounded-grid amplifier. (a) Through r-f chokes; (b) through special low-capacity transformer (for a-c only); (c) through quarter-wavelength line; (d) through coaxial choke.

destroyed. Similarly, if the tube has a suppressor grid, it too must be returned to the ground. In lower-power tubes, the suppressor grid is often returned to the cathode internally. In this case, the tube is not usable for grounded-grid circuits, at least not at very high frequencies. The a-c grounding of all the grids makes a multigrid triode out of the tube, since each grid is being driven by the input signal. The screen grid still must be placed at the proper d-c voltage in accordance with the class of operation and the manufacturer's data.

3-13. The Grounded-plate Amplifier. *General Characteristics.* The grounded-plate amplifier, more commonly known as the *cathode follower*, is seldom used in transmitter power-amplifier r-f circuitry, although it has found use as a driver power amplifier in cases where the r-f load may vary. A typical application has been as a driver for a grounded-grid television r-f power amplifier in which the load presented to the driver varies with the video modulation of the radio frequency.

A practical difficulty in applying the cathode follower is the special care necessary to prevent overdriving the tube. The grid-to-cathode r-f driving voltage is a small percentage of the total r-f voltage existing from grid to ground. Therefore small misadjustments in the tuning of either the input or output tuned circuits can easily result in overdriving the tube and hence result in excessive grid dissipation.

The cathode follower has a relatively large amount of coupling between the input and output tuned circuits due to the close spacing between the control grid and cathode. Thus the cathode follower will require neutralization at all but the very lowest frequencies. As shown in Fig. 3-16, the cathode follower with a tuned circuit in the grid, also forms the basic circuit of a Colpitts oscillator and will often behave as such.

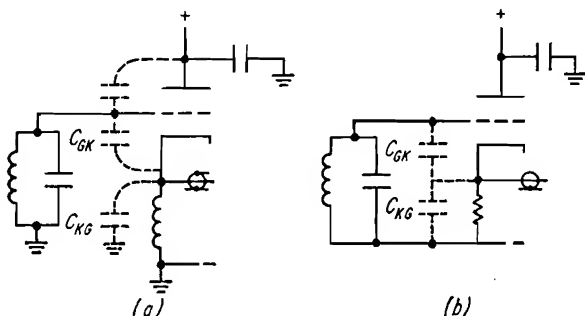


FIG. 3-16. A cathode follower with a tuned circuit in its grid forms a Colpitts oscillator because of stray capacities as shown. (a) Cathode-follower amplifier; (b) Colpitts oscillator equivalent.

The stray capacities that exist between the grid and cathode and from cathode to ground form the necessary feedback path for the oscillator.

In general, the cathode follower is characterized by a higher input impedance, lower output impedance, and less than unity voltage gain. However, since its use in transmitter r-f circuitry is rather limited, a more detailed discussion of this tube configuration is better left to other texts.

3-14. Amplifier Output Load Resistance. An important consideration is the amplifier load resistance into which the amplifier must operate. This figure is inherent in the amplifier design methods previously outlined for the various types of amplifier. The output load resistance of any amplifier is related to the amplifier plate-voltage swing and the fundamental r-f component of the plate current. Since these values were determined as part of the amplifier design, they can be used to determine the amplifier load resistance by the following:

$$R_L = \frac{E_{pm}}{I_{pm}} = \frac{E_{bb} - E_{b\min}}{I_{pm}} \quad (3-17)$$

If the values required by the above equation are unknown, a good approximation of the load resistance is given by

$$R_L \simeq 0.5 \frac{E_{bb}}{I_b} \quad (3-18)$$

This figure is important in the design of the output matching network between the amplifier and the load (which is normally the antenna). The design of various matching networks is discussed fully in Chap. 5. Any departure of the output matching network from the desired impedance transformation characteristic will be reflected in the performance of the amplifier. If the load impedance seen by the amplifier is lower than originally desired, then the plate-voltage swing will not be as great (for the same drive or excitation) and the efficiency will be poorer. If the resistance is higher than the original design value, then the power output will be lower, and the efficiency will, in all probability, be somewhat higher.

3-15. Amplifier Stabilization. Most amplifiers require some care and consideration in their design and construction to prevent spurious oscillations from occurring in the amplifier and associated circuitry. These oscillations take three general forms in a transmitter, each of which require different methods of suppression.

The first oscillation is that which may occur at approximately the same frequency as the amplifier operating frequency. This is due to either or both of two possible reasons. The first is feedback through the coupling capacitance of the amplifier tube. This results in the amplifier acting like a tuned-plate-tuned-grid oscillator. The cure for this oscillation is to cancel the feedback capacitance by means of a neutralizing circuit, or to use a tetrode, or in some cases to use both. The second cause of oscillation at the amplifier operating frequency is through direct external feedback between the tuned input circuit of the amplifier and the output circuit. This can take place through the inductive field of the coils that are part of the input and output networks, or capacitively through the placement of other components associated with the input and output circuits. The cure for this is obvious; but the cure is often not easy, particularly at the vhf region and if the amplifier is already built. The initial placement of parts is a prime consideration. Shielding should be provided where necessary. The bypassing and decoupling between power supplies should also be carefully considered.

The second form of oscillation that may occur in an amplifier is due to stray capacitances and inductances in the grid and plate circuits resonating and forming an oscillator in the vhf region. The feedback path may be through the tube or through some common impedance at the oscillat-

ing frequency. This is usually cured by inserting a high impedance at the oscillating frequency in the feedback path in the grid and plate.

The third oscillation that may occur is a low-frequency one due to the r-f choke in the grid and plate resonating with the stray capacities present in the circuit and forming a tuned-plate-tuned-grid oscillator.

Detection of Amplifier Self-oscillation. Any instability in an amplifier can usually be detected by operating the amplifier without any r-f excitation and without any fixed bias. In order not to exceed the plate and screen dissipation ratings of the tube, it is also usually necessary to reduce the operating voltages to reduce the dissipation. Now, by observing the plate, screen, and grid current meters on the amplifier as the plate and grid tuning controls are adjusted throughout their ranges, any tendency toward oscillation can be observed on the meters as a change in the current readings. The control-grid current should read zero under all circumstances. Any control-grid current flow indicates some form of r-f oscillation. The resulting oscillation can be any of the forms noted previously. The method of cure is determined by the frequency and the cause of the oscillation. A grid-dip meter or wavemeter can be used to both detect and determine the frequency of the oscillation. If the frequency is lower than the frequency range of the wavemeter or grid-dip meter being used, then the harmonics can be observed. The difference between any successive harmonics observed on the wavemeter or grid-dip meter is the frequency of the oscillation. A more sensitive detector to use is a receiver. This would be desirable as a final check, after the amplifier is thought to be completely suppressed.

VHF Parasitic Oscillation in Amplifiers. Spurious or parasitic oscillations are quite common in new amplifier designs unless precautions have been taken to suppress the tendency toward any vhf oscillations. The general frequency of the vhf parasitic oscillation is 40 to 150 Mc in the typical tetrode amplifier. The parasitic oscillation is the result of the stray capacities and inductances connecting the amplifier components together. These form resonant circuits at the vhf frequencies. Furthermore, the screen grid usually loses some of its effectiveness at the vhf frequencies owing to the construction of the tube. The internal structure within the tube becomes an appreciable impedance at the vhf region, making the screen ineffective. Further difficulty is encountered in making the screen-grid bypassing effective at the vhf parasitic frequency. Thus the coupling between the input and output through the tube together with the stray inductances and capacitances present will allow the tube to perform as a tuned-plate-tuned-grid oscillator in the vhf region as well as an amplifier at the lower frequencies.

This spurious oscillation requires power to sustain it, and therefore it reduces the power output of the amplifier by just that amount. If the

amplifier is to operate in the linear mode, then a self-oscillation of this sort will tend to change the amplifier characteristics. Unfortunately some parasitic oscillations occur only at modulation peaks, or under certain conditions of tuning. Therefore care should be exercised in ascertaining if the parasitic is truly suppressed after suppressing the more obvious parasitic oscillations.

Suppressing parasitic oscillations can be more of an art than a science, but there is no magic involved in successfully achieving results. It is difficult to predict the feedback path of the vhf oscillatory circuit, but this is usually not necessary, for all but the more stubborn cases can be eliminated through the use of parasitic suppressors and through careful and thorough bypassing for the vhf frequencies. The typical parasitic suppressor consists of four to ten turns of wire wound around a 50- to 100-ohm noninductive resistor. These suppressors are placed in the amplifier circuit in series with the grid and plate leads and preferably directly at the grid and plate terminals of the tube. The exact number of turns required for any particular amplifier design is determined by the physical layout of the components and the type of tube used, and therefore must be determined experimentally. For amplifiers operating in the h-f region and lower, the presence of a parasitic suppressor in the plate and grid leads will not effect the performance of the amplifier. However the impedance offered by the suppressor will effectively break up the vhf oscillatory circuit that may be formed. The resistor across the parasitic choke may be sometimes successfully eliminated. This is a convenience in higher-power amplifier where this resistor may become quite large, since in higher-power amplifiers the parasitic power is likewise increased. The damping of this oscillation or even the tendency toward oscillation will require a resistor with an adequately large power-dissipation capability.

Although the amplifier is designed for a much lower frequency than the parasitic frequency, the bypassing capacitors in the amplifier must be good for both the parasitic frequency and the operating frequency. The internal inductance of a capacitor, which may be entirely satisfactory at the lower operating frequency, may be entirely inadequate at the vhf parasitic frequency. Therefore, the capacitor which was intended to present a low impedance now presents a high impedance to the vhf parasitic oscillation. The resultant common impedance will also provide a feedback path and can result in oscillatory condition.

The successful elimination of parasitic oscillation thus requires four general steps.

1. On the day the amplifier's physical layout is determined, short leads should be provided among the components to break up the multiplicity of vhf resonances that can occur.

2. The bypassing and coupling capacitors are chosen to ensure that the impedance of the capacitor is not appreciable at the potential vhf oscillatory frequencies and that no vhf resonance exists in the components selected.

3. The amplifier is tested for any spurious output frequencies by the means previously described.

4. These parasitic frequencies are suppressed by the means also previously discussed, and rechecked even more carefully to be sure that they have been completely suppressed, i.e., not merely moved in frequency or partially suppressed, only to reappear on modulation peaks, etc.

Amplifier Neutralization. A completely neutralized amplifier must fulfill two conditions. The first is that the interelectrode capacitance

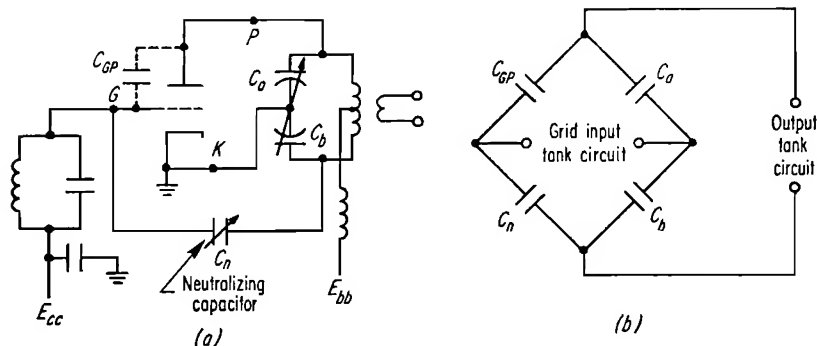


FIG. 3-17. A neutralized amplifier and the equivalent bridge neutralization circuit. (a) Typical triode neutralized amplifier; (b) equivalent bridge circuit.

between the input and output circuits be canceled. The second requirement is that the inductance of the grid and cathode assemblies and leads be completely canceled. Cancellation of these common impedances between the input and output will theoretically prevent oscillation. This, of course also applies in practice as well but often not without some difficulty.

There are a variety of methods of accomplishing these ends that will fulfill the two conditions. In amplifiers operating up to the high frequencies it is not necessary to cancel the impedances of the grid and cathode leads. However the interelectrode capacitances cannot be ignored, particularly in amplifiers using high-gain tubes. As the gain of the amplifier increases, the need to cancel the feedback voltage becomes that much more necessary. For this reason, it is usually necessary to neutralize tetrodes at the higher frequencies.

Many systems of neutralization are based on forming a balanced bridge with the tube grid-to-plate capacity forming one of the arms of the bridge. For example, Fig. 3-17a which illustrates a typical system,

can be redrawn to the equivalent bridge circuit illustrated in Fig. 3-17b. From this circuit, we can see that, if the bridge is balanced, there will be complete isolation between the input circuit and the output circuit.

The bridge circuit shown in Fig. 3-17 is inherently capable of remaining in balance at all frequencies. This would infer that the stage, once neutralized, is neutralized at all frequencies. Unfortunately this is not always true. In practice the variations, in lead lengths in the arms of the bridge make the problem of achieving wideband neutralization more critical. If these stray inductances are essentially equal, then wideband

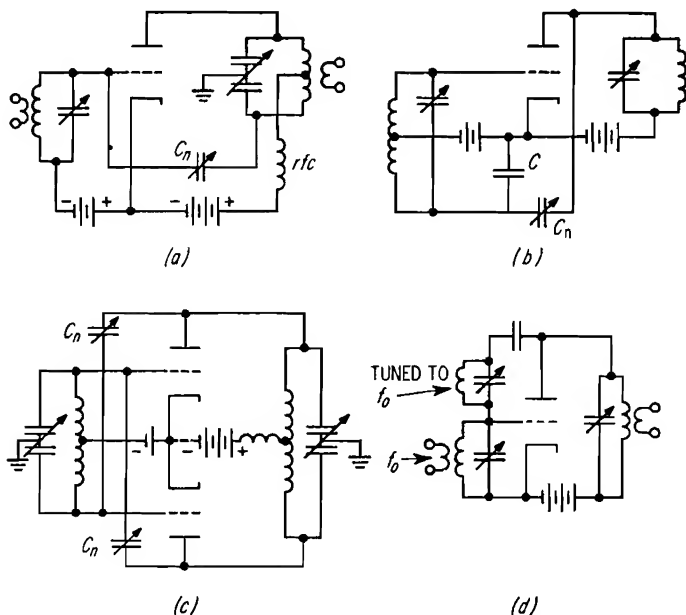


FIG. 3-18. Typical methods of neutralizing amplifier. (a) Plate neutralization (Hazeltine system); (b) grid neutralization (Rice system); (c) push-pull or cross neutralization; (d) coil neutralization.

neutralization can be achieved. Other various systems of bridge-neutralizing triodes are shown in Fig. 3-18a, b, and c.

Figure 3-18d illustrates another method of effectively canceling the feedback path from grid to plate. This is done by making the grid-to-plate capacitance part of an unloaded parallel-tuned circuit. The unloaded impedance of this circuit is very high compared to the loaded impedance of both the plate- and grid-tuned circuits, and therefore the coupling between the plate and grid under these circumstances is very small. This method requires that the resonant frequency of the grid-plate-tuned circuit be readjusted for any changes in the amplifier frequency. However this is hardly any disadvantage for fixed-station use.

“Neutralization of Tetrodes (17). If neutralization on the fundamental frequency below the vhf region is found desirable, normal cross-neutralization of push-pull amplifiers is simplest. For tetrodes the neutralizing capacitors are small, and each capacitor need only be a wire connected to each side of the grid circuit, brought through the chassis deck, and allowed to ‘look’ at the plate of the tube on the opposite half of the circuit. The wire or rod can be $\frac{1}{2}$ to 1 in. away from the glass, and, by adjusting its length or spacing, the last trace of coupling can be eliminated from the amplifier. A simple adjusting collet mounted on the amplifier chassis will support the wire or rod and allow it to be adjusted.

“In the case of the single-ended stage, either a push-pull output or a push-pull input circuit can be used to provide the out-of-phase voltage necessary for neutralization. Because of the low voltage and the small

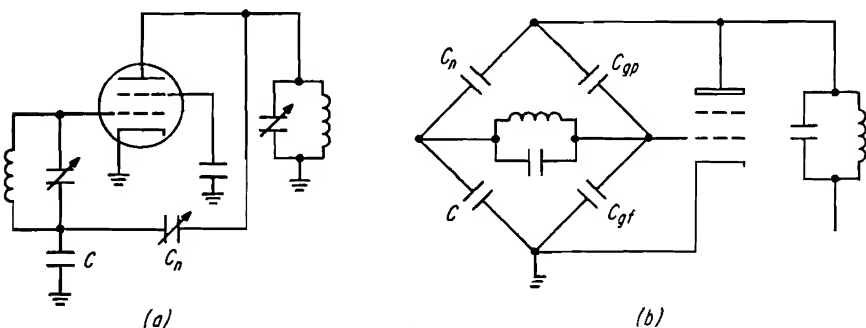


FIG. 3-19. Single-ended grid neutralization. (a) Basic method for neutralizing a tetrode; (b) equivalent bridge circuit.

size of the input r-f circuit, it is usually simpler to make the input circuit push-pull, and the circuit becomes a grid-neutralized circuit (see Fig. 3-18b). The neutralizing capacitor C_n is again very small and similar to those described above. To maintain the balance of the input circuit while tuning, it is desirable to have a padding capacitor C_i equal in size to the input capacitance of the tetrode.

“This system is suitable up to the vhf region. However this method assumes that the screen is well bypassed to ground and so provides the expected screening action. In the vhf and uhf regions the screen is not necessarily at r-f ground potential, and the neutralizing becomes more involved.

“Single-ended r-f stages can also be easily grid-neutralized without using the conventional push-pull input circuit (see Fig. 3-19). In this method described by Bruene the input resonant circuit is taken off ground by a small amount by making the input-circuit bypass capacitor C somewhat smaller than usual. The voltage to ground across capacitor C

is out of phase with the grid voltage and can be fed to the plate to provide neutralization. In this case the neutralizing capacitor C_n is considerable larger than the capacitance from plate to grid, and is about the size of those used for neutralizing triodes.

"The basic circuit of the Bruene method is shown in Fig. 3-19a. It can be redrawn as a capacitance bridge showing clearly the grid-neutralization circuit (see Fig. 3-19b). Balance is achieved when

$$\frac{C_n}{C} = \frac{C_{gp}}{C_{gf}} \quad (3-19)$$

where C_{gp} = feedback capacitance grid to plate of tetrode, and C_{gf} = total input capacitance including tube and stray capacitance.

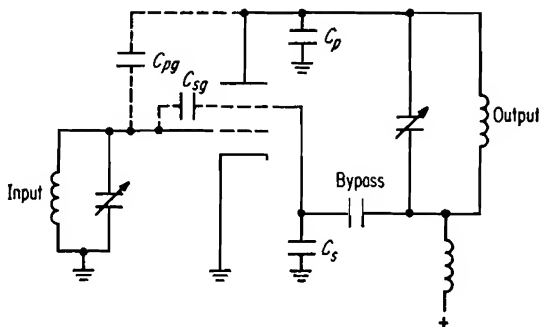


FIG. 3-20. Basic circuit of single-ended neutralization of a tetrode.

"Single-ended amplifiers can also be neutralized by taking the plate circuit a small amount off ground, as was done in the single-ended grid neutralizing scheme, and by using the tube capacitances as part of the bridge circuit. This method has been described by Hultberg.

"Such a circuit is shown in Fig. 3-20. It differs from the usual r-f amplifier circuit in that the plate bypass capacitor is returned to the screen side of the screen bypass capacitor C_s and the stray capacitance from the plate to ground C_p is added. The size of the screen bypass capacitor C_s and the amount of stray capacitance in C_p are chosen to balance out the voltages induced in the input circuit by the internal tube capacitances to grid C_{pg} and screen to grid C_{gs} .

"The circuit is redrawn in Fig. 3-21 in the usual bridge circuit form. Balance is achieved when

$$\frac{C_p}{C_s} = \frac{C_{pg}}{C_{gs}} \quad (3-20)$$

In usual tetrode structures the capacitance from the screen to grid is roughly half the published tube input capacitance. (The tube input

capacitance is mainly the sum of the capacitances of the grid to screen and the grid to cathode. Since as a first guess these two capacitances are roughly equal, one is not far off in using half the listed tube input capacitance.)

"It should be noted that in all neutralizing capacitance bridge circuits it is assumed that the frequency is low enough so that the inductances

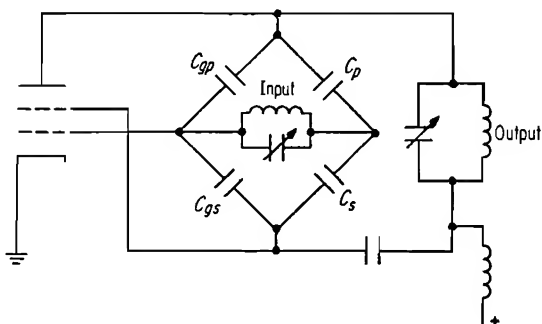


FIG. 3-21. Single-ended plate neutralization showing capacitance bridge circuit that is present.

in the connecting leads and the tube structures can be neglected. In the vhf and uhf regions this is often not true. The voltage developed across this residual inductance is often used to accomplish the neutralization at these frequencies.

"Careful analysis of the feedback circuits of tetrodes in the vhf region has been made, and the basic concepts follow. In Fig. 3-22 the tetrode-

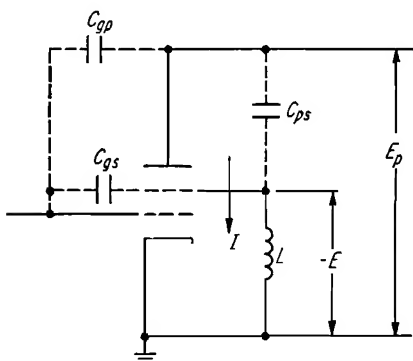


FIG. 3-22. Feedback paths for a tetrode amplifier.

in the screen leads. The passage of this current through the inductance L develops a voltage $-E$ which has a polarity opposite to that of the plate voltage E_p .

circuit elements involved in the feedback circuits are indicated. These circuit elements are inherent, inside the vacuum enclosure of the tube, and involve the residual capacitance from plate to grid, the capacitance from plate to screen, the capacitance from screen to grid, and the inductance of the screen lead of the tube. It will be noted that the r-f voltage developed in the plate circuit E_p causes a current I to flow through the plate to screen C_{ps} , and the inductance L

"In Fig. 3-23 these same circuit elements and voltages have been arranged with a graphical representation where the height above or below the zero line represents magnitude and polarity of the r-f voltage, of that part of the circuit with respect to zero or filament voltage. Because all the circuit components involved are pure reactances, the voltages are either in phase or out of phase, and so can be represented as positive and negative with respect to each other. The voltages plotted are the components only of the r-f output circuit voltage E_p , and no attempt is made

to show the normal driving voltage on the grid. The plate P is shown at a high positive potential above zero, and the magnitude is represented by the distance above the zero line as shown by the dimension E_p . The voltage developed in the screen lead inductance places the screen at a negative voltage with respect to the plate voltage. The screen of the tube S is shown to be below the filament line, or negative, by the amount $-E$. If the circuit were perfectly neutralized, the control grid G would lie on the

zero potential line or at filament potential in so far as any action of the r-f plate voltage E on the input circuit is concerned. If there is no component of output voltage developed between grid and filament, the circuit is not neutralized.

"The total r-f voltage between plate and screen comprises plate voltage E_p and the screen lead inductance voltage $-E$. This total voltage is applied across a potential divider consisting of the capacitance plate to grid C_{pg} in series with the capacitance grid to screen C_{gs} . When this potential divider is suitably matched to the magnitudes of the voltage E_p and the screen lead voltage $-E$, the control grid will have no voltage difference to filament as a result of the output circuit voltage E_p .

"It should be noted in Fig. 3-23 that the potential dividing action between capacitances plate to grid C_{pg} and grid to screen C_{gs} will not be affected by the operating frequency. It should be noted also that the division of voltage between plate and screen and screen and ground due to the charging current I will vary greatly with frequency. There will, therefore, be some particular frequency at which this potential dividing circuit places the grid at filament potential as far as the plate circuit action is concerned, and this is called the *self-neutralizing frequency* of the tetrode. At this particular frequency, the tetrode is inherently

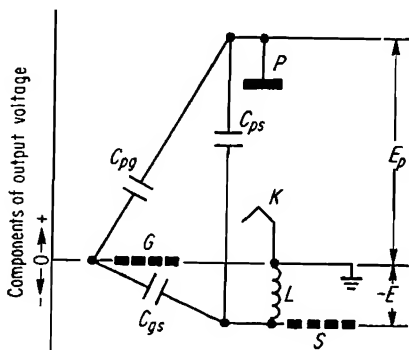


Fig. 3-23. Graphical presentation of voltages in self-neutralized tetrode. (Courtesy of Eitel-McCullough, Inc.)

neutralized due to the circuit elements within the tube structure and any external added screen lead inductance to ground. Typical self-neutralizing frequencies with normal screen bypassing circuits are as given in Table 3-4.

TABLE 3-4. APPROXIMATE SELF-NEUTRALIZING FREQUENCIES OF TETRODES

<i>Tube Number</i>	<i>Frequency, Mc</i>
4-1000A.....	25-30
4-400A.....	45
4-250A.....	45
4X500F.....	75-90
4-125A.....	75-90
4-65A.....	80-120
4X150A }	Apparently above the useful range of the tube
4X150G }	
4X500A }	

"When the tube is operated below the self-neutralizing frequency, the normal cross-neutralizing circuits apply. In this case a neutralizing capacitor approximately equal to the plate-grid capacitance of the tube brings voltage of opposite polarity from the output circuit to the grid or from the input circuit to the plate.

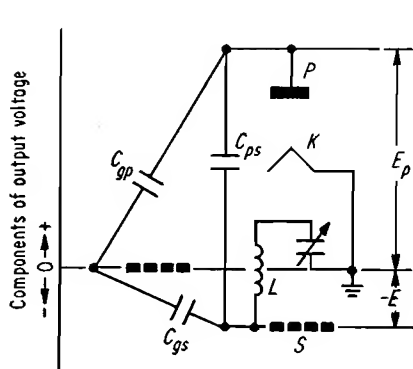


FIG. 3-24. Graphical presentation of voltages in tetrode with added screen-lead capacitance. (Courtesy of Eitel-McCullough, Inc.)

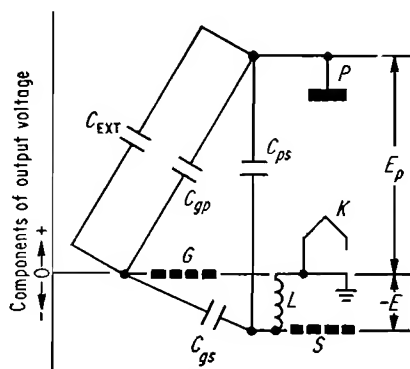


FIG. 3-25. Graphical presentation of voltages in tetrode neutralized by added external grid-to-plate capacitance. (Courtesy of Eitel-McCullough, Inc.)

"If the operating frequency is higher than the self-neutralizing frequency of the tetrode, the voltage $-E$ developed in the screen-grid lead inductance is too large to give the proper voltage division between the internal capacitances of the tube. One obvious method of reducing the voltage in the screen-lead inductance is to series-tune the screen lead to ground so as to lower the total reactance. This takes the form of a series variable capacitor as shown in the graphical representation in Fig. 3-24.

"Another method would be to change the potential divider network made up of the tube capacitances. This could be done by adding capacitance external to the tube between grid and plate. The method is shown in Fig. 3-25. This added capacitance plate to grid is on the same order of size as the residual grid-plate capacitance of the tetrode and hence is similar in construction to the neutralizing capacitance used at lower frequencies. However, in this case the small wire or rod standing up beside the tube "looking" at the plate (and so forming a neutralizing capacitor) is connected to the grid of the tube rather than to an opposite polarity in the input circuit.

"If the r-f power amplifier, operating above the self-neutralizing frequency of the tube, must tune over a range of frequencies, it is probably easier to use the screen series-tuning capacitor method and make this control available to the operator. If operation is desired over a range of frequencies including the self-neutralizing frequency of the tube, this circuit is also desirable because the incidental lead inductance in the variable tuning capacitor lowers the self-neutralizing frequency of the circuit so that the neutralizing series capacitor can be made to operate over the total desired frequency range. Obviously, if the range is too great, switching of neutralizing circuits will be required. Usually, a small 50- or 100- μf variable capacitor in the screen lead has been found satisfactory."

Neutralizing the Grounded-grid Amplifier. Since the grid acts as a shield between the input and output circuits of a grounded-grid amplifier, the need for neutralizing this type of amplifier is often avoided. While it is entirely possible to neutralize the plate-to-cathode capacitance present in the grounded-grid amplifier by the same general means generally employed in the grounded-cathode amplifier previously discussed, usually it is neither necessary nor desirable. The addition of plate-to-cathode neutralizing capacitors increases the minimum shunt capacitance. In vhf circuitry, this is most undesirable as previously pointed out. The low output capacitance is one of the main features of the grounded-grid amplifier.

A convenient method of stabilizing the grounded-grid amplifier is simply by making the grid-to-ground path slightly inductive. The resultant positive feedback thus introduced cancels the feedback current fed through the plate-to-cathode capacitance. In practice, the inductance within the tube assembly combined with the grid connecting structure will provide this inductive path. In some cases, this inductance is too great, resulting in oscillation. A series capacitor will cancel a portion of the inductive reactance to bring about the neutralized condition described.

General Amplifier-neutralization Procedure. The procedure for neutralizing any amplifier generally follows the same procedure as that out-

lined here. However specific pieces of equipment may have a particular procedure recommended by the manufacturer, which procedure should be followed for reasons peculiar to that piece of equipment.

The first step is to remove the d-c plate voltage from the amplifier if the tube is a triode. If the amplifier tube is a tetrode, then both the plate and screen voltage should be removed. In either case the r-f circuit should not be broken. The d-c connection can either be disconnected physically, or the plate or plate-screen voltage can be merely turned off. A sensitive r-f meter can then be coupled to the amplifier output circuit, and the neutralizing adjustments made for the minimum

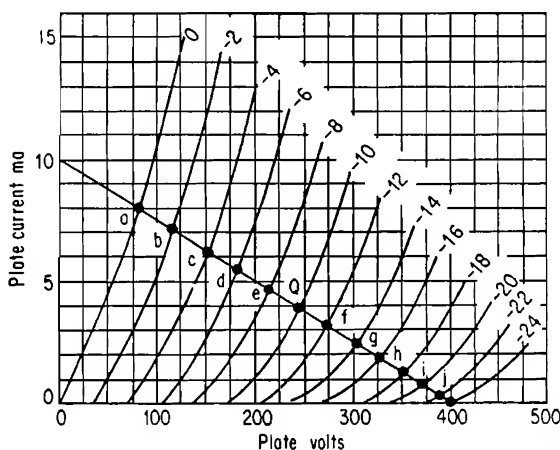


FIG. 3-26. Typical load line on triode characteristic curve. (From Thomas L. Martin, Jr., "Electronic Circuits," Prentice-Hall, Inc., Englewood Cliffs, N.J., 1955.)

r-f output. The amplifier-grid current meter can also serve as a neutralization indicator simply by observing the reaction of the grid current as the plate circuit tuning capacitor is rocked back and forth through resonance while the neutralization control is being adjusted. Neutralization is achieved when the reaction of the plate tuning on the grid current is minimized.

A final neutralizing adjustment is often required at the higher frequencies. This can be done by operating the amplifier at the fully loaded normal condition and observing the plate-current dip and the grid current peak as the plate circuit is rotated through resonance. For correct neutralization the grid-current peak and the plate-circuit dip should occur simultaneously. Care should be observed that the amplifier plate dissipation is not exceeded when it is turned off resonance.

3-16. The Linear Amplifier. General. A linear amplifier is simply one in which the output voltage is directly proportional to the input voltage. In other words, the transfer characteristic for a linear amplifier

is a straight line when plotted on a graph of input vs. output. From this definition, not all types and classes of amplifiers can be described as linear. From the previous descriptions of amplifier classes and their operation we can see that the class A amplifier can be made to operate in a linear manner.

Linear operation is not inherent in the class A amplifier but is a characteristic that is designed into the amplifier circuit by proper selection of the operating conditions. For example, Fig. 3-26 illustrates a typical triode characteristic tube curve, with a load line designating the operation of the amplifier. From this curve, the dynamic transfer characteristic curve can be constructed by replotting the input voltage vs. the output plate current. The output plate current is directly proportional to the output voltage. Thus the replotted curve will actually be the input vs. output voltage, or the transfer characteristic curve of the amplifier. In this case, the transfer curve just obtained and illustrated in Fig. 3-27 has a slight curve, indicating some degree of non-linearity. By operating the tube within the definition of the class A amplifier with different load lines, i.e., different plate voltages, load resistances, and bias voltages, the transfer characteristic can be made more or less linear. Actually no amplifier is without some degree of distortion which indicates some degree of nonlinearity. However the degree of distortion that can be tolerated in a linear amplifier is determined by the application and the system requirements.

The class B amplifier can also be considered as a linear amplifier with the same qualifications as the class A amplifier given above but with one additional qualification. Specifically, a single tube operated as a class B amplifier must have a resonant-tuned circuit in its output in order to be capable of linear operation. Since the class B amplifier is biased to cutoff, it produces no plate current when the input a-c or r-f signal swings in the negative direction. If the load were a resistor, and if no plate current were flowing, there would be no output voltage. Under these circumstances, the class B amplifier is not linear, since no output appears for one-half of the input cycle even though there is an input signal being applied. However the same amplifier with a tuned circuit as part

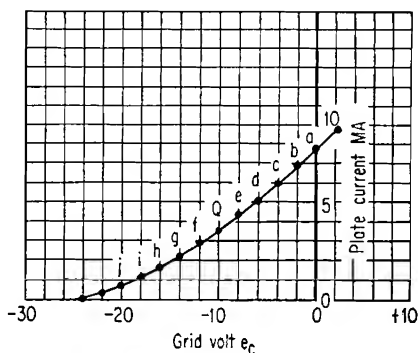


FIG. 3-27. Transfer characteristic replot from Fig. 3-26. (From Thomas L. Martin, Jr., "Electronic Circuits," Prentice-Hall, Inc., Englewood Cliffs, N.J., 1955.)

of the load resistance will behave quite differently. The resonant tuned circuit will supply the missing half cycle not supplied by the tube. Thus a single-ended tuned class B amplifier can be made to operate in a linear manner. A push-pull class B amplifier can be operated linearly with just a resistance load since each tube will alternately supply one-half of the a-c output signal.

In general, both the class A and class B amplifiers, together with their combinations such as the class AB₁ and AB₂ amplifiers, can all be operated as linear amplifiers.

The class C amplifier must, by definition, operate with the plate current flowing for less than 180° in the input cycle. This requires that the tube be biased beyond plate-current cutoff. Thus, for a-c input signals that do not exceed the difference between the grid-bias voltage and the tube cutoff grid voltage, there will be no output of any kind from the amplifier. Since there is an input signal being applied, with no resultant output, the amplifier must be considered as nonlinear. For this reason, the class C amplifier is basically not capable of operating linearly and still remain within the classification.

Efficiency Characteristics of the Linear Amplifier. The fact that an amplifier is linear infers that it must handle some form of modulated r-f signal. This signal may be a voice- or video-modulated amplitude modulated, r-f carrier, or a single-sideband signal. The particular type of driving signal will determine the average efficiency of the amplifier. This average efficiency is the average of the instantaneous efficiencies of the amplifier taken over the modulation cycle. For example, if the driving signal to a linear amplifier is an r-f signal amplitude-modulated by a steady sinusoidal tone, say 1,000 cps, as shown in Fig. 3-28a, the efficiency of the amplifier at the peak of the modulating cycle might be 70 per cent. However, the efficiency over the entire modulation cycle would average out to only 35 per cent. If this same amplifier were to handle an ssb signal, also modulated at a steady 1,000 cps, the efficiency of the amplifier would be a constant 70 per cent. This is because of the nature of an ssb signal. A steady tone is actually a steady or continuous r-f signal as illustrated in Fig. 3-28b. In this latter case, the amplifier can operate at full peak capacity. However, if the input r-f signal drops for either of the two cases just discussed, the amplifier efficiency will decrease proportionately.

The dependency of the amplifier on the character of the driving signal might best be pursued a little further by considering the conditions in a linear amplifier with an a-m signal applied to the amplifier input. In a linear amplifier the plate voltage is fixed. Therefore with an increase in the r-f drive voltage, the only thing that can vary is the plate current. For example, if the drive signal voltage doubles, then the plate current

should double. Under these circumstances, the power input to the amplifier has also doubled. However, the fundamental fact exists that under these exact circumstances the power output must have increased four times. This quadruple increase in the power output, despite the fact that the power input has only doubled, is accounted for by the increased amplifier efficiency at the higher drive levels. This increase in efficiency is directly proportional to the efficiency at the peak. For example, if the peak efficiency is 70 per cent, at half-drive level the

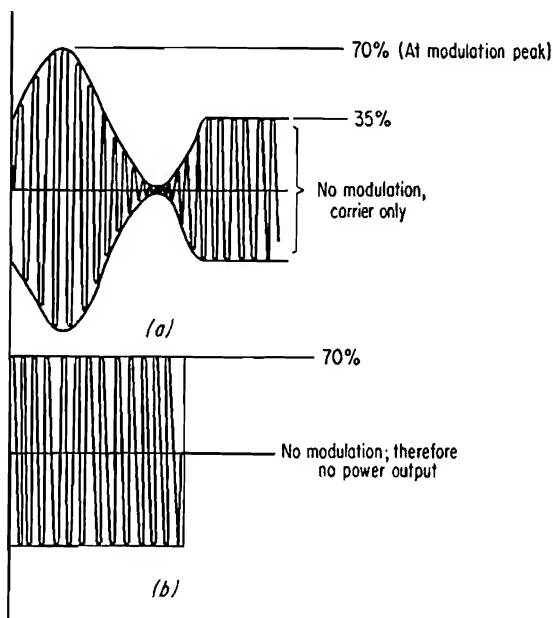


FIG. 3-28. Comparison of efficiency for amplifier handling an a-m and ssb signal. (a) Sine-wave modulated a-m signal; (b) sine-wave modulated ssb signal.

efficiency is 35 per cent; at one-quarter-level, the efficiency would be 17.5 per cent, etc.

Thus if the amplifier is used for voice-modulated a-m service, the average efficiency will be quite poor. Normally it will not exceed 35 per cent for class B operation, and it will be proportionately less as the amplifier operation approaches class A. Although the average efficiency of the linear amplifier operated with an a-m signal for drive is quite low, it should not be immediately discounted in favor of other systems with what might appear to be higher efficiencies. Efficiency must be considered from the over-all transmitter viewpoint. A linear amplifier operated as a class AB₁ amplifier requires very little drive power. Thus, in terms of over-all transmitter efficiency, there is no real advantage in the typical class C amplifier with its separate high-

power modulator and higher-power driver stages. In many cases of high-power a-m transmitters there is little choice, for it becomes very expensive and somewhat difficult to design and construct a high-power modulation transformer for the tens of kilowatts. It is usually more economical to buy this in the form of a higher-power a-c power supply to supply the less efficient linear amplifier.

The linear amplifier finds a great deal of use as a power amplifier for television transmitters. In these cases, the driving r-f signal is a video-modulated a-m signal. Since the video information contained in the modulation contains a d-c component which corresponds to the picture brightness, the level of the efficiency of the amplifier operated in this service will vary according to the scene brightness and the video used to modulate the signal. For example, with an amplifier operating at black level, i.e., with no video modulation other than the synchronizing pulses, the amplifier efficiency will be the highest. With the amplifier operating at white level, i.e., with the scene brightness corresponding to white and with no video modulation, the amplifier efficiency will be the poorest. This is in accordance with the previous discussion on amplifier efficiency and its dependency on the drive signal.

Distortion in the Linear Amplifier. Perhaps the most important characteristic of a linear amplifier is its distortion level. The distortion present in an amplifier is a measure of the linearity of the transfer characteristic and, in general, a measure of how well the amplifier is doing as a linear amplifier. The tolerable distortion level depends on the service requirements of the amplifier. Actually, all amplifiers exhibit some degree of nonlinearity and hence will exhibit some degree of distortion. The degree of distortion can be minimized by special treatment and care in the design. Special attention must be given to the selection of the operating conditions, i.e., to the plate and bias voltages and also to the regulation of the power supplies and the r-f driver.

Any nonlinearity in an amplifier produces distortion products which are the result of the sum and difference combinations of the harmonics of the r-f frequencies being fed into the amplifier. Since the power amplifier will normally use some form of tank circuit in the output, the distortion products which are not within the bandpass of the tank circuit are filtered out and will not appear in the output. However those distortion products which lie close to the original frequencies and are within the bandpass of the tuned circuits of the amplifier will appear in the output. These distortion products are more commonly known as *odd-order* distortion products, and more specifically as the third-order, fifth-order, seventh-order, etc., distortion products.

The third-order distortion products correspond to $2f_1 - f_2$ and $2f_2 - f_1$, where f_1 and f_2 are any two r-f signals present in the original signals.

Similarly, the fifth-order product is $3f_2 - 2f_1$ and $3f_1 - 2f_2$. The output spectrum of an amplifier showing the possible distortion products is illustrated in Fig. 3-29.

In wideband amplifiers handling many telephone channels, odd-order distortion can result in interference in other channels outside the one producing the interference. In both narrow-band and wideband amplifiers these distortion products can cause interference outside of the bandwidth being used to transmit the intelligence, since the r-f bandwidth of most output tank circuits is considerably wider than the intelligence being transmitted.

It is apparent that since distortion is of prime concern in linear amplifiers, it would be of interest to ascertain the causes and cures for nonlinearities. As illustrated previously in Fig. 3-27, even the class A amplifier characteristic exhibits some curvature. The distortion can be kept to acceptable levels in this case by operating the amplifier at low signal levels and by operating the signal at the most linear portion of the transfer characteristic.

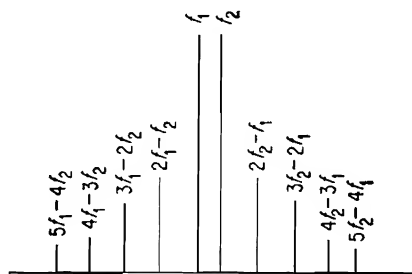


FIG. 3-29. Spectrum of odd-order distortion products in an amplifier with input frequencies of f_1 and f_2 .

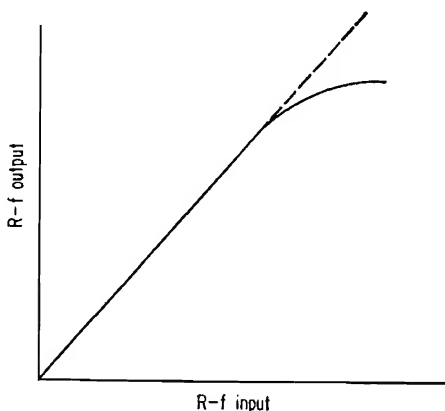


FIG. 3-30. The effect of grid loading on amplifier linearity.

In the class AB amplifiers, the static plate current is made as high as is acceptable within the limits of the tube plate dissipation, the available power-supply ratings, and the distortion that can be tolerated.

Another common cause of nonlinearity is the result of driving the amplifier into the grid-current region. This causes the linearity to drop at the larger signal levels as shown in Fig. 3-30. A secondary effect of

driving the amplifier into the grid-current region is the effect of the varying r-f load presented by the amplifier to its driver. Since the inherent regulation of most amplifiers is poor, the effect of the varying load on the driver amplifier will result in a nonlinear transfer characteristic. A common way of minimizing this effect is to use swamping resistors at the driver amplifier output as shown in Fig. 3-31. To achieve good regulation, these resistors are arranged to absorb approximately ten times the power delivered to the grid. Thus the

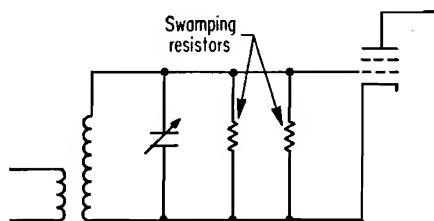


FIG. 3-31. The effect of amplifier grid current on linearity can be minimized by loading the input circuit with swamping resistors.

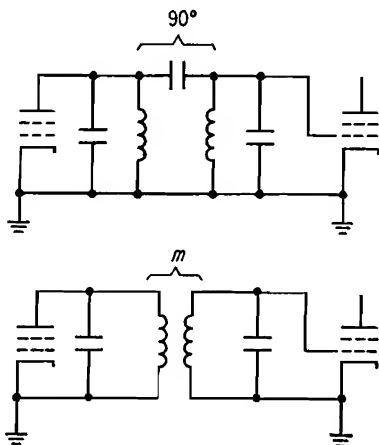


FIG. 3-32. Two basic impedance-inverting networks to provide low driving impedance from high-impedance drivers. (D-c circuits not shown for simplicity.)

effect of the varying grid power is minimized on the driver. The effect on the driver output with a varying load can be minimized in another way through the use of an impedance-inverting network as shown in Fig. 3-32. However this type of circuit has the disadvantage of requiring careful adjustment which becomes difficult if the transmitter is intended for general coverage.

Perhaps the best way to eliminate the problem of driver regulation is through the selection of amplifier tubes which do not require drive power to develop the required power output, i.e., tubes which can operate and develop power in class AB₁.

A general requirement in all linear amplifiers other than class A is good power-supply regulation. Once again the value of the regulation depends on the amount of distortion that can be tolerated. For example, when large signals are fed through a linear amplifier, the resulting large plate swings will put a heavier current demand on the screen supply (since the plate voltage will swing to values around or even lower than the screen voltage). If the screen voltage drops at this time, the plate current and power output will drop, resulting in some nonlinearity.

The distortion introduced by an amplifier or a whole series of cascaded amplifier stages can be reduced still further by means of feedback. This

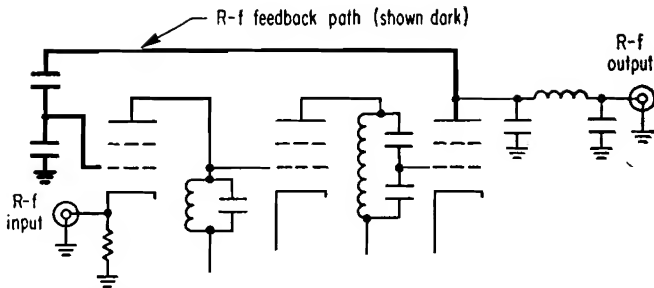


FIG. 3-33. Basic r-f feedback method for reducing distortion in linear amplifier chain.

feedback can be a direct r-f feedback as shown in Fig. 3-33, or it can be an envelope feedback as shown in Fig. 3-34. In general, the r-f feedback method is the superior method, although they are both used quite successfully in commercial equipments. For signal to distortion ratios much beyond 30 db, the use of feedback becomes increasingly attractive as a means of achieving the desired distortion level. Ten decibels of r-f feedback will generally increase the signal-to-distortion ratio of an amplifier by an additional 10 db.

Methods of Checking Amplifier Linearity.

Distortion, or nonlinearity, can be checked by the simple circuit illustrated in Fig. 3-35. The detectors shown in this circuit should have the same r-f level applied to the detector to ensure that both crystals are operated under the same

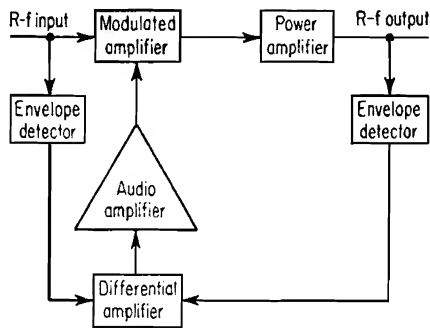


FIG. 3-34. Basic method of envelope feedback for reducing distortion in linear amplifiers.

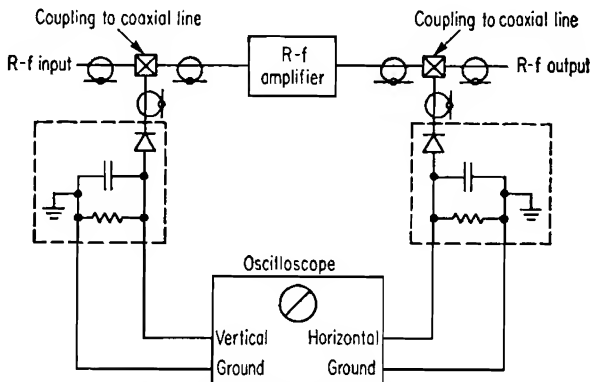


FIG. 3-35. Basic method of checking amplifier linearity.

dynamic conditions. For best results, the crystals should be matched; however, for ordinary applications this is usually not necessary. The detector can be checked for match simply by interchanging the assemblies and noting the change in the oscilloscope pattern. If the detectors are the same, then there will be no change in the presentation on the

oscilloscope. The degree of nonlinearity of the amplifier is illustrated on the scope face by the amount of the curvature of the line. A truly linear amplifier will present a perfectly straight line on the oscilloscope. Typical oscilloscope patterns are shown in Fig. 3-36. This method is not suitable for observing or detecting very low levels of distortion but is useful for initial tune-up adjustments of linear amplifiers.

A second more sensitive method utilizes a receiver to search for the odd-order distortion products. This requires that the amplifier be driven by two r-f signals of slightly different frequencies. The receiver must have the selectivity required to select the third-order distortion signals. The level of the distortion can then be measured by means of the calibrated output attenuator on a signal generator if the absolute value of the distortion is desired. For relative readings of distortion, the *S* meter on the receiver itself will suffice.

A third method of checking the distortion in a linear amplifier is known as the two-tone test. This method is particularly applicable to linear amplifiers used in single-sideband applications. For this reason, a detailed discussion on this method is reserved for Chap. 6 on single sideband.

3-17. Wideband Amplifiers. *General.* Amplifiers intended for wideband service find their greatest use in the vhf region and higher where the spectrum space is available. Television and multichannel telephone systems form the bulk of these applications. In the majority of these applications, the wideband power amplifier is also required to be linear. Therefore special considerations must be given the selection of the operating conditions of such an amplifier.

In the wideband amplifier, particular attention is given to the load impedance into which the tube is operated. The lower the load imped-

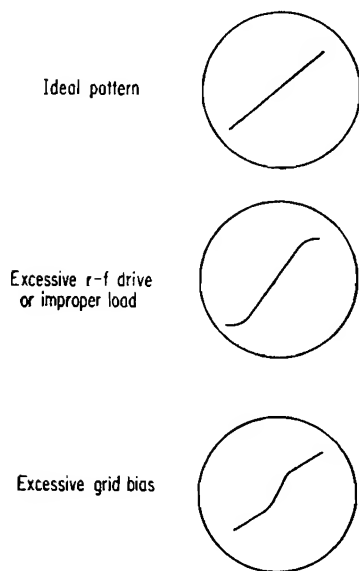


FIG. 3-36. Typical waveforms produced by circuit shown in Fig. 3-35.

ance, the lower will be the power output for a given plate current. The load impedance is determined mainly by the bandwidth of the output tuned circuits and the output capacitance of the amplifier. Specifically, the load impedance is inversely proportional to both the bandwidth and output capacity. Therefore, to obtain the maximum power output for a given bandwidth, it is necessary to minimize the output capacity of the amplifier. The output capacity is largely dependent on the tube itself (its size and construction), the arrangement of the components associated with the output circuit (which in turn results in stray capacities), and lastly on the amplifier configuration. As pointed out previously, the unneutralized grounded grid exhibits the lowest output capacity. The output capacity of this configuration is essentially the capacity existing between the plate and the grounded control grid or screen grid. For this reason it is the usual choice in this type of operation.

Once the choice of tube and circuit has been established, and once the bandwidth (between 3-db points) has been firmly established, the only way the load impedance into which the tube will operate may be varied is by the choice of output circuit. A simple single-tuned output circuit will result in the lowest load impedance, and hence the least power output. The load impedance for a single-tuned circuit is given by

$$R_L = \frac{1}{2\pi\Delta fC} \quad (3-21)$$

A double-tuned circuit will present a slightly higher load impedance to the tube. For a transitionally coupled tuned circuit, i.e., one in which the coupling has been adjusted for maximum flatness, the load impedance is given by

$$R_L = \frac{1}{\sqrt{2}\pi\Delta fC} \quad (3-22)$$

Thus we can see that the double-tuned circuit presents a load resistance 1.414 times greater than the single-tuned circuit. Of course the double-tuned circuit is considerably more difficult to tune and adjust and is more difficult and expensive to construct initially. Actually this load impedance can be increased even further by using triple-tuned circuits; but because of the practical difficulty in adjusting and maintaining these circuits, it is not commonly used in transmitting equipment. Ladder network output circuits also show promise of eventual use for wideband use but at present are not used to any great extent.

Wideband-amplifier Design. If it is desired to determine the operating point of a wideband class B linear amplifier using a double-tuned output circuit, the first step would be to determine the output capacity from the

available tubes, based on the power level desired. Proper allowance should be made for the inevitable stray capacities.

For example, suppose it is desired to develop 25 kw in a TV amplifier. This power rating is that developed at the peak of the synchronizing pulse. From a survey of the available tubes at this power level, the type 4W20,000 is selected. This tube has an output capacity of 23 μmf . Allowing 10 μmf for stray capacity, the total output capacity is 33 μmf . The output load resistance as seen by the tube would then be found by substituting in Eq. (3-22) as follows:

$$R_L = \frac{1}{\sqrt{2} \pi \Delta f C} = \frac{1}{\sqrt{2} \pi (10 \times 10^6) (33 \times 10^{-12})} = 680 \text{ ohms} \quad (3-23)$$

The alternating current flowing through this load at the peak power level is then

$$I_{\text{rms}} = \sqrt{\frac{P}{R_L}} = \sqrt{\frac{25,000}{680}} = 6.07 \text{ amp} \quad (3-24)$$

In a class B amplifier the peak plate current is

$$I_{\text{peak}} = 2 \sqrt{2} (I_{\text{rms}}) = 2 \sqrt{2} (6.07) = 17.2 \text{ amp} \quad (3-25)$$

This value together with the minimum value of the plate-voltage swing to be determined will be point 1 on the load line as described in Sec. 3-10 on amplifier design. The minimum value of the plate-voltage swing is determined simply by choosing a point along the peak plate-current line calculated in the step above before the line starts to rise sharply.

The required value of the plate voltage is

$$e_{bm} + E_{pm} = E_{bb} \quad (3-26)$$

This value together with the bias voltage forms point 2 of the load line. Points 1 and 2 can be connected together and the design completed by the means described in Sec. 3-10.

3-18. Frequency Multipliers. In the majority of transmitters, the frequency generator operates at a much lower frequency than the desired transmitter output. The frequency multiplication to the desired output frequency is accomplished by means of amplifiers variously called *frequency multipliers* or *harmonic generators*. These amplifiers are basically class C amplifiers with the output tank circuit tuned to a multiple or harmonic of the input frequency. Since the class C amplifier plate current flows in pulses, these pulses will contain a high percentage of harmonics. In the normal straight-through amplifier, the harmonics present in the current pulses are filtered out by the tank circuit. However, if the tank circuit is tuned to the desired harmonic, then, of course, the harmonic will be favored.

The angle of plate current flow is important if optimum output is desired from a given multiplier stage. In general, the narrower angles favor the higher harmonic frequencies. Table 3-5 gives a summary of the plate current pulse length and power output for frequency multipliers. The data of Table 3-5 represent a practical compromise between two conflicting considerations in the optimizing of frequency-multiplier design. If the plate-current conduction angle is narrowed, the r-f drive power will increase since the bias voltage must be increased. At the same time the plate efficiency increases. However, the same limiting factors still exist as discussed previously in Sec. 3-9 where the limiting factor in increasing the plate efficiency is the fact that the power output rapidly decreases. If the plate-current pulse is made longer, the pulse will have less harmonic content available to excite the output tank circuit.

The design of a frequency multiplier stage follows the same general procedure described earlier for a class C amplifier, except that the plate current conduction angle is chosen in accordance with the figures presented in Table 3-5. Similarly the adjustment is identical with that of the class C amplifier, except that the output tank circuit is tuned to the harmonic rather than the fundamental input frequency.

TABLE 3-5. PLATE-CURRENT PULSE LENGTH AND POWER OUTPUT OF HARMONIC GENERATORS*

Harmonic	Optimum length of pulse, electrical degrees at the fundamental frequency	Approximate power output, assuming that the normal class C output is 1.0	Relative load impedance, assuming that class C case is 1.0 (approximately)
2nd	90-120	0.50-0.65	1.5-2.0
3rd	80-120	0.30-0.40	2.5-3.3
4th	70-90	0.25-0.30	3.3-4.0
5th	60-72	0.20-0.25	4.0-5.0

* F. E. Terman, "Electronic and Radio Engineering," p. 475, McGraw-Hill Book Company, Inc., New York, 1955.

Frequency-multiplier Circuits. The circuit for a single-tube frequency multiplier is identical to that for a standard straight-through amplifier. The difference in operation is accomplished by electrical adjustment of the amplifier operating conditions.

A frequency multiplier using a push-pull amplifier circuit is capable of output at the odd harmonics of the input frequency only. The push-pull configuration cancels the even harmonics because of the balanced nature of the circuit.

The push-push multiplier will produce output at even harmonics only. It will not produce any output at the fundamental or odd harmonics. This circuit is shown in Fig. 3-37. A unique feature of the push-push arrangement is that if one of the tube filaments is turned off, the circuit functions as a straight-through amplifier. The disabled tube acts as a neutralizing capacitor, preventing oscillation. Normally, a frequency multiplier does not require neutralization since the plate and grid are tuned to different frequencies. However, with high transconductance tubes, the circuit may oscillate as a tuned-plate-tuned-grid oscillator. In these cases neutralization will be required. The same precautions are necessary to prevent a frequency multiplier from parasitic oscillations as were taken in the straight-through amplifier.

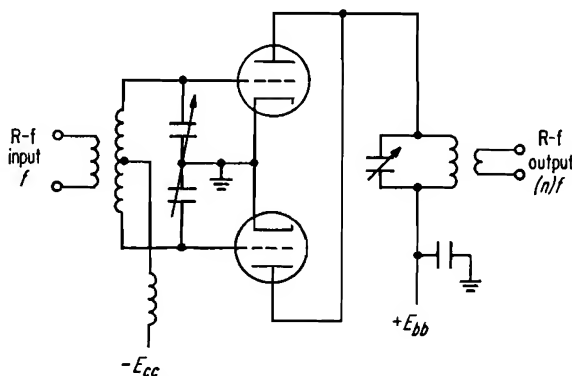


FIG. 3-37. Basic circuit of a push-push frequency multiplier.

3-19. Practical Circuit Considerations. *Amplifier Bias.* All negative grid amplifiers must have their control grids fixed at some negative bias voltage determined by the amplifier design considerations specified in preceding sections. There are three general methods of obtaining this negative grid-bias voltage, namely through the use of a fixed-bias power supply, by means of a self-bias resistor, and through the use of a combination of fixed bias and self-bias.

Fixed bias of an amplifier utilizes a separate power source which develops a fixed voltage as illustrated in Fig. 3-38a. Since the grid will draw current when normal r-f drive is applied (assuming the amplifier is operating with grid current) this grid current will pass through the supply. This grid current will tend to produce a voltage across the internal resistance of the grid supply which will increase the negative grid bias still further. Thus if a purely fixed bias of the grid is desired, the bias supply must have a low internal resistance. If the supply is regulated, the grid supply voltage will remain constant with or without

grid current flowing in the amplifier. (Assuming the power supply is subjected to conditions within its range of regulation.) Fixed bias is used particularly in applications involving linear amplifiers for TV where the r-f drive varies according to the modulation, and where a power level refers to a specific brightness level. Fixed bias is also necessary in grid-modulated r-f power amplifiers.

Self-bias, sometimes called resistor bias, depends on the r-f drive to develop the negative grid bias for the tube. This is useful only in

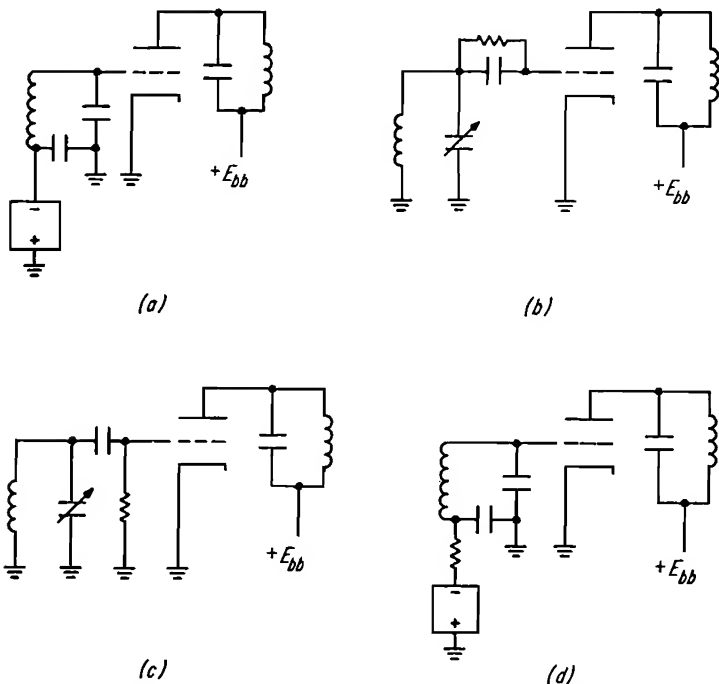


FIG. 3-38. Basic methods of amplifier grid biasing. (a) Fixed bias; (b) series self-bias; (c) shunt self-bias; (d) combination of self and fixed bias.

applications in which the drive level is essentially constant. Particular care must be used in applying fixed bias to an amplifier since if the r-f drive fails, the bias voltage will drop to zero. In most cases the plate dissipation of the amplifier will be exceeded if some precautionary steps are not taken.

In low-level applications, a cathode resistor can be used to limit the plate current and hence the plate dissipation in case the bias voltage disappears. In higher-power tetrode applications, a screen clamper circuit, as shown in Fig. 3-39, can be used to limit the plate dissipation in the event the r-f drive fails. If the bias drops to zero, the clamper tube begins to conduct and will drop the screen-grid voltage to a low value.

The resistance value of the self-bias resistor is found simply by applying Ohm's law. The d-c grid-bias voltage divided by the d-c grid current will give the required value of resistance. Both the d-c grid voltage and current can be determined from previously described amplifier calculations or from the manufacturer's literature.

The third general method of grid biasing is to combine the two methods just described. Some amount of fixed bias is used to limit the plate current in the event the r-f drive fails, and the remaining value of bias voltage is made up with self-bias. The self-bias can be developed across a resistor in series with a low-internal-resistance bias supply. If the bias supply has a high internal resistance, the grid current can develop

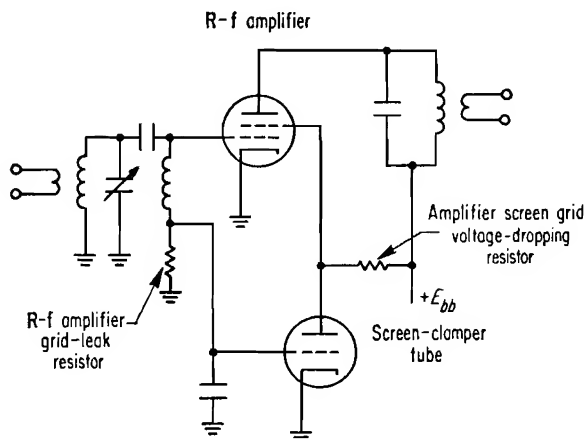


FIG. 3-39. Basic circuit for screen clamping to protect tetrode r-f amplifier tube from loss of excitation.

the required self-bias across the bleeder resistance of the fixed-bias supply. Some of these methods are shown in Fig. 3-38.

Screen-grid Supply. The screen grid in the tetrode can be supplied either through a separate power supply or through a dropping resistor from the plate-voltage power supply. The choice of the method used to supply the screen grid in a tetrode or an amplifier is determined by the amplifier application. A separate fixed-voltage screen-grid supply is necessary for all linear r-f amplifiers and grid-modulated r-f amplifiers for much the same reason that fixed grid bias was found to be necessary for these applications. If the r-f excitation is lost in a tetrode amplifier and if the screen voltage were not fixed in these applications, then any change in the screen current would also result in a change in the screen voltage. This would result in a change in the operation conditions of the amplifier from the original specified design conditions. In any case, the use of a fixed screen supply with a fixed grid-bias supply provides

complete protection against exceeding the plate dissipation due to loss of r-f excitation.

A fixed screen-grid supply used with a self-biased r-f amplifier tends to reduce the effect of variations in the r-f excitation of the amplifier. This limiting effect or saturating effect may be desirable in many cases.

A screen-grid dropping resistor is usually used in a plate-modulated r-f amplifier since it is necessary for the screen grid to be modulated along with the plate to achieve 100 per cent modulation of a tetrode tube. If a fixed screen-grid supply were used in this application, a special modulation transformer would be required to modulate both the plate and the screen grid separately. If a screen-grid dropping resistor is used with a self-biased amplifier, it is necessary to use a system of amplifier tube protection previously shown in Fig. 3-39. If the excitation were lost with just a screen-grid dropping resistor alone, then the screen-grid current drops to practically zero, in which case the screen voltage would rise to practically the plate voltage. Since the control-grid cutoff voltage of the screen-grid voltage is approximately one-fifth of the screen-grid voltage, the tube conditions will change rapidly, and the plate dissipation rises sharply.

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POWER TUBES

4-1. Transmitting-tube Construction. *General.* Tubes for transmitting purposes differ only slightly in principle from those used for receiving purposes in the same frequency range. In low-power applications the two types are used interchangeably. However, as the power level increases, the physical changes become more obvious. The increased power-handling capabilities usually result in increased size even at the uhf range. Higher power generally requires higher voltages to achieve the higher currents. The higher voltages require more spacing between the tube elements to prevent arc-over between tube elements. This tends to increase the size of the tube. This then requires careful and unique construction techniques to minimize lead inductance from the internal tube element to the circuit. Increased power requires more electron emission from the filament or cathode, which also increases the size and heat generated by this element. In addition, higher power also requires increased heat-dissipation capabilities in the grid and anode structures. Adequate cooling means must then be provided for the tube. This is normally achieved by either air convection currents, forced-air cooling, or by water or liquid cooling.

4-2. The Cathode. *General.* The emitting material for the cathode in transmitting tubes is generally either pure tungsten, thoriated tungsten, or a mixture of the oxides of calcium, barium, and strontium. These materials, to varying degrees, exhibit good emission characteristics at a reasonable cathode temperature. Higher cathode temperatures increase the heat-dissipation problems in the tube. The cathode material should also exhibit a minimum evaporation of the emitting material. Excessive evaporation would tend to deposit the emitting material on the grid which might result in emission directly from the grid. The resulting grid-current flow is in the opposite direction to the normal grid-current flow and in some cases may cause an actual reversal of the grid-current meter.

The cathode should also be relatively resistant to damage from positive ion bombardment. This results from the electrons emitted from the cathode striking random gas atoms present in the tube. The electrons

will knock some electrons from the atom, forming a relatively heavy positive ion that will be attracted to the more negatively charged cathode.

Tungsten Cathodes. Tungsten cathodes are widely used in high-voltage tubes. It is characterized by its long life, its high mechanical strength, and its resistance to damage by positive ion bombardment. This latter characteristic results from the fact that the emission characteristics do not depend on any surface film of emitting material as in other types of cathodes. Thus, if the surface of the cathode is damaged by positive-ion bombardment, there is no effect on the emission of the cathode.

However, tungsten requires a high temperature for satisfactory emission. The working temperature of tungsten is generally in the order of 2550°K , which results in a dazzling white cathode color in normal operation. At this temperature, the evaporation rate is still sufficiently low, which ensures long cathode life.

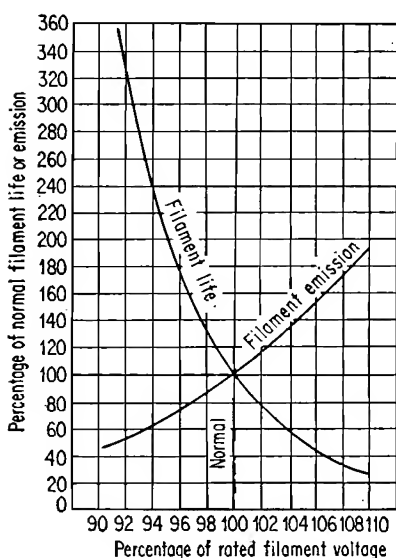


FIG. 4-1. Effect of filament voltage on the life and emission of a tungsten filament.

As a result of the high working temperature of tungsten, the power required to heat the cathode to this working temperature is higher than other cathode materials. At a cathode temperature of 2550°K the specific emission (milliamperes of emission per watt of heating power) is 7 ma per watt.

Tungsten Cathode Life. The theoretical life of the tungsten cathode depends on the rate of evaporation of the tungsten. Since the tungsten cathode is not dependent on any surface file for emission, the cathode may be operated at complete saturation; i.e., all of the electrons that are emitted are attracted to the plate with no space charge being formed. Thus, if the circuit in which a particular tube using a tungsten cathode

does not demand the full peak emission from the tube at that particular working temperature, the working temperature of the cathode may be lowered to produce less emission. The tube may be operated in this manner indefinitely with no damage. The chart of Fig. 4-1 illustrates the relatively theoretical life expectancies and emission characteristics obtained by various filament voltages for a tungsten filamentary cathode.

It should be emphasized that adjustment of the cathode temperature to extend tube life applies only to the tungsten cathode.

Thoriated Tungsten Cathodes. The specific emission of the cathode can be increased to a higher level at a lower working temperature by the addition of approximately 2 per cent thorium to the tungsten. At a cathode temperature of 2000°K (bright yellow) the specific emission of a thoriated tungsten cathode is between 70 and 100 ma per watt.

The source of emission in this type of cathode is a thin layer (one atom thick) of thorium driven to the surface of the cathode by an initial "flashing" process performed by the manufacturer. In operation, as the thorium evaporates, it is replaced by thorium driven to the cathode surface by the cathode temperature. In order to maintain a balance between the evaporation and replacement of the thorium at the surface of the cathode, it is necessary to maintain a narrow range of cathode temperatures. Increasing the cathode temperature drives more thorium to the surface. However, the surface layer of thorium will remain only one atom thick. The excess thorium driven to the surface is evaporated, which shortens the life of the tube. The correct cathode temperature is obtained by following the manufacturer's recommendations for the filament voltage and current. Either raising or lowering the filament voltage from these values will only result in a shortened tube life.

If a thoriated tungsten filament loses its emission characteristics, owing to either a severe overload or low-filament-voltage operation, it can sometimes be rejuvenated by operating the filament at its normally rated voltage for 10 min or more with no other voltages applied to the tube. This process can be accelerated by increasing the filament voltage 20 per cent above normal for a few minutes. If normal filament activity is not restored by this process, then the following treatment may be attempted. Operate the filament at 70 per cent above the normal operating filament voltage for 5 min. This can then be reduced to 20 per cent overvoltage for a period of 15 min. The plate voltage should be removed during this treatment.

The Oxide-coated Cathode. This type of cathode has the highest specific emission of all the types of cathodes discussed. It is generally formed from the carbonates of barium, calcium, and strontium, which are converted to the oxides of these elements during the tube processing. Specific emissions of 200 to 300 ma per watt at working temperatures of 1060°K (dull red color) are common. Because the emitting material is applied to a nickel cylinder or directly to a filament, the surface can be damaged by positive-ion bombardment. This is minimized by maintaining the emission from the cathode at a much higher level than that required by the peak current through the tube. This forms a space charge which cushions the bombardment by the positive ions. If the

anode current becomes limited by the emission of the cathode, the oxide coating may actually be pulled off the cathode. Like the thoriated tungsten cathode, the oxide-coated cathode temperature must be carefully maintained between narrow limits in order to obtain the maximum life.

Generally, oxide-coated cathodes are found in tubes with plate dissipation under 100 watts. However, some higher-power tubes have been developed using oxide-coated cathodes, and further use of these cathodes may be made in the high-power range.

Cathode Construction. There are two general types of cathodes in common use, i.e., the directly heated cathode and the indirectly heated cathode. In the directly heated or filamentary cathode, either the emitting material is coated on an electrically heated filament, or the filament itself is made of an emitting material such as tungsten.

In the indirectly heated cathodes, the filament is entirely separated and insulated from the emitting material of the cathode. The advantage of this arrangement is that the cathode does not carry any filament current, and therefore all parts of the cathode will be at the same d-c potential. This condition does not exist in the filamentary type of cathode, and it often results in hum modulation of the r-f output if some precautions are not taken. Hum can be mini-

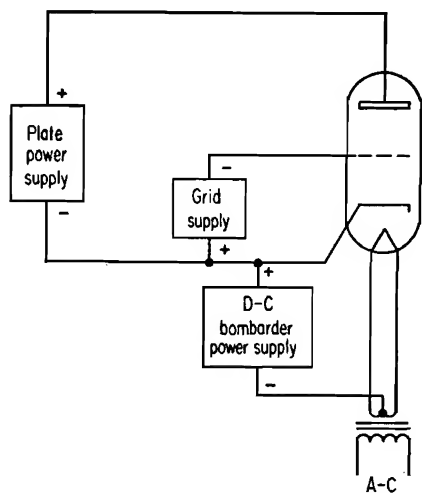


FIG. 4-2. D-c circuit of a tube with a bombarder cathode.

imized by the use of center-tapped transformers for the filament supply in which the center tap is used for the d-c return lead to the power supplies. A second alternative is simply to run the filament from a d-c power supply.

In many higher-powered tubes, the shape of the cathode makes it difficult to heat by conventional indirect methods. The specific cathode shape may be required by the high-frequency design of the tube, or the shape may be convenient for use in an electron gun assembly for use in a klystron, etc. In some of these cases, an indirectly heated cathode arrangement such as shown in Fig. 4-2 is used. The cathode is actually the anode of a diode while the filament functions as the cathode of the diode. A high voltage placed across the filament and cathode will cause the cathode (or diode anode) to heat up to emitting temperature. This arrangement is known as the *bombarder* type of cathode.

4-3. Grids. Since the grid is in the immediate vicinity of the filament, it must be able to dissipate the heat absorbed from the filament. The amount of heat generated by the grid itself is negligible compared to that absorbed through radiation from the filament. For this reason, the metal employed by the grid is related to the type of cathode employed, the power rating of the tube, and the proximity of the grid to the cathode. The metals commonly used for the grid are tungsten, molybdenum, and tantalum, and nickel-manganese alloys.

The heat can be conducted from the grid by means of a heavy grid-support structure and dissipated out through the leads, or it can be radiated by means of black cooling fins attached to the grid supports. The heat absorbed by the grid can be minimized by making the grid a good reflector of heat. Gold- or platinum-plated grids are often used with oxide-cathode cathodes. These are not only good reflectors, but since these metals do not absorb oxygen from any of the oxides that may be deposited on the grid from the cathode, they minimize any secondary emission from the grid. Platinum plating of the grid is generally preferred over gold since it can withstand higher temperatures.

The control grid in many classes of tube operation is positive with respect to the cathode, and in the case of the tetrode the screen grid is often more positive than the anode. Under these conditions, electrons attracted to the grids may hit with force sufficient to liberate electrons from the grid. This "secondary emission" can be minimized by coating the grid with a carbon film.

4-4. Anodes. General. The construction and material used for the anode in a vacuum tube will depend on the anode dissipation rating and the method used for dissipating the heat absorbed by the anode. The three commonly used methods of cooling the anode are (1) by the flow of normal air convection currents, (2) by forced-air cooling, (3) by liquid cooling. Convection cooling is usually in only the lower-power applications where the anode is enclosed in the glass envelope. Heat absorbed by the anode is radiated through the vacuum to the envelope where the air convection currents will remove the heat. In the cases of forced-air and liquid cooling, in all but the lowest-power applications, the anodes are of the external type, i.e., the external part of the anode is in direct contact with the cooling air or liquid.

There is a fourth method of cooling the anode in vacuum tubes, this one known as *evaporative cooling*. It is very effective and efficient; however, it is not used to any large extent at present. In this process, the coolant liquid placed in contact with the anode is allowed to vaporize. In this way, a smaller amount of coolant will remove a larger amount of heat than by any other means.

Convection-cooled Enclosed Anodes. In low-power tubes the anode is generally made of carbonized nickel or nickel-plated steel. The car-

bonizing process gives the surface a black finish which functions as a more efficient radiator than a bright surface. This permits the anode to run at a lower temperature and reduces the temperature of the nearby grids. It also reduces the possibility of grid emission due to excessive temperatures.

Higher-power types of enclosed anode tubes use tungsten, molybdenum, graphite, and tantalum for the anode material. Tungsten anodes will maintain their shape under high temperatures. They are also relatively free of gas and are easily degassed. However, tungsten is so difficult to work into the necessary shape required by most anode structures that it is not used very often in modern tubes.

Molybdenum also degasses readily and is much more easily formed into the shapes required by the anode structure. Its disadvantage is its low thermal emissivity which is partially overcome by roughening the surface with a carborundum blasting and by the use of fins to increase the radiation area of the anode.

Tantalum is similar to molybdenum except that its melting point is slightly higher and is more easily worked. It does have the added advantage of absorbing stray gases released from the various tube elements during tube operation.

Graphite anodes are quite satisfactory in most respects. They inherently contain more gas which can be removed during the tube manufacture. However, they do not warp, and because of the thickness of the graphite anode, hot spots are minimized. As a result, graphite anodes tend to run cooler for the same dissipation compared to metal anode tubes. Graphite also provides a gettering action which absorbs stray gases.

Zirconium is often coated on the surface of metallic anodes. This metal also provides a gettering action which absorbs any stray gases released in the tube during operation, thus maintaining a high vacuum in the tube.

External Anodes. External anodes can be cooled by either air or water. In either case, the material typically used for this purpose is oxygen-free copper. Copper is easily machined and has both a high electrical and heat conductivity. Although it has a lower melting temperature compared to the materials used in the radiation-cooled enclosed anode tubes, the external anode tube rarely exceeds an average anode temperature of 200°F.

Forced-air-cooled external anodes typically use a fin cooler. This is comprised of a number of copper fins silver-soldered to a copper hub which is part of the anode of the tube. This presents a large radiating surface to the airstream forced through the cooler. This is necessary since a forced-air-cooled tube will typically dissipate only 3 to 4 watts

per sq in. at 150 to 200°C. For comparison, a liquid-cooled anode will dissipate from 200 to 700 watts per sq in. of anode surface with temperatures between 30 and 70°C. The dissipation per square inch of cooler area can, of course, be increased simply by increasing the velocity of the air through the cooler. However, the increased air noise that results generally limits this rating.

The water-cooling jacket surrounding the liquid-cooled tube may be either an integral part of the tube, or it may be removable. The trend in modern tube practice is toward the integral jacket type. This type has advantages when it is desired to mount the anode end up, in a circuit. However, the removable jacket permits the anode scale to be easily removed.

4-5. General Types of Transmitting Tubes. Transmitting tubes can be broadly classified according to very general types, i.e., the negative-grid type and the velocity-modulated types. The negative-grid classification applies to all the commonly found triodes, tetrodes, pentodes, etc., in which the electron passage to the anode is controlled by a negatively biased grid on which a signal is impressed. The detailed operation of the negative-grid tube has been described in such detail in so many books that it is not necessary to go into further detail in the operation of this type of tube.

The second general type of tube uses the principle of velocity modulation of an electron beam to achieve the amplification. This type of tube finds wide application in the vhf range and higher frequencies in the form of klystrons or traveling-wave tubes. The basic operation of this type of tube will be described in succeeding pages of this chapter.

4-6. Klystron Amplifier Tubes. General. The klystron amplifier finds wide general use in communications and radar equipment at frequencies above 250 Mc. At these frequencies, the klystron has an advantage in that the r-f circuitry is independent of the cathode and anode and so has none of the problems associated with transit time of the electrons between the cathode and anode as in the more conventional grid-controlled tubes. The klystron also has the added advantage in that the r-f circuitry is at ground potential. Thus there is no problem with bypass capacitors and d-c insulation.

Basically, the klystron consists of an electron gun, a means of focusing the electron beam through a small *drift* tube, some r-f cavities interspersed at intervals along the drift tube, and a collector or anode. The means of focusing is usually magnetic, although electrostatic focusing is used on some types. A cross section of a klystron amplifier is shown in Fig. 4-3.

The number of cavities in a klystron determines the bandwidth or gain capabilities of the amplifier. The more cavities in the klystron, the wider will be the bandwidth, or the greater will be the power gain of the

amplifier. Tubes with power gains of up to 90 db have been made, but gains in the order of 35 db for a typical three-cavity tube are more typical. Tubes with up to six cavities have been made for wideband applications.

Klystron Amplifier Operation. The klystron is one of the group of tubes that operates on the principle of velocity modulation of an electron beam. The electron gun for the klystron of Fig. 4-3 is comprised of the

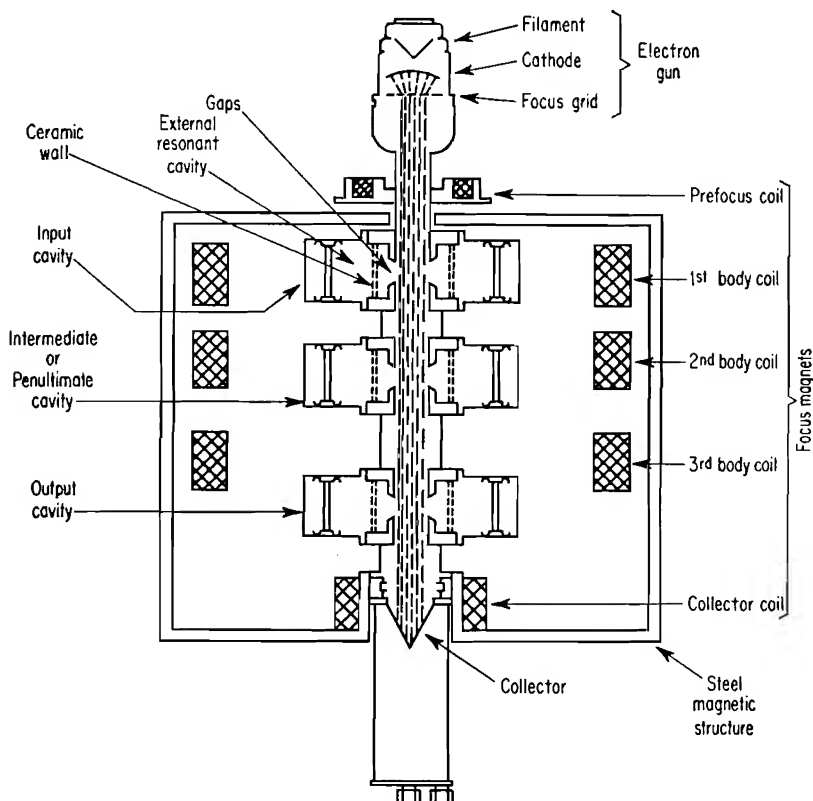


FIG. 4-3. Cross section of typical klystron.

cathode, the focus electrode, and the prefocus magnet. In this tube, the cathode is an oxide-coated concave section of a sphere. As electrons are emitted toward the collector or anode, they pass the focus electrode. This electrode is either at the cathode potential or is at a slightly negative potential with respect to the cathode. This forces the beam to the negatively charged focus electrode, and the negatively charged electrons repel each other, thus forcing the beam to converge as it enters the drift tube. A small prefocus magnet acts like a magnetic lens to direct and concentrate the beam before it enters the main magnetic field of the klystron.

An axial magnetic field is required throughout the length of the klystron because the electrons comprising the beam mutually repel one another. This would tend to spread the beam. The axial magnetic field surrounding the entire drift-tube length focuses or concentrates the electron beam as it travels towards the collector. The individual magnet

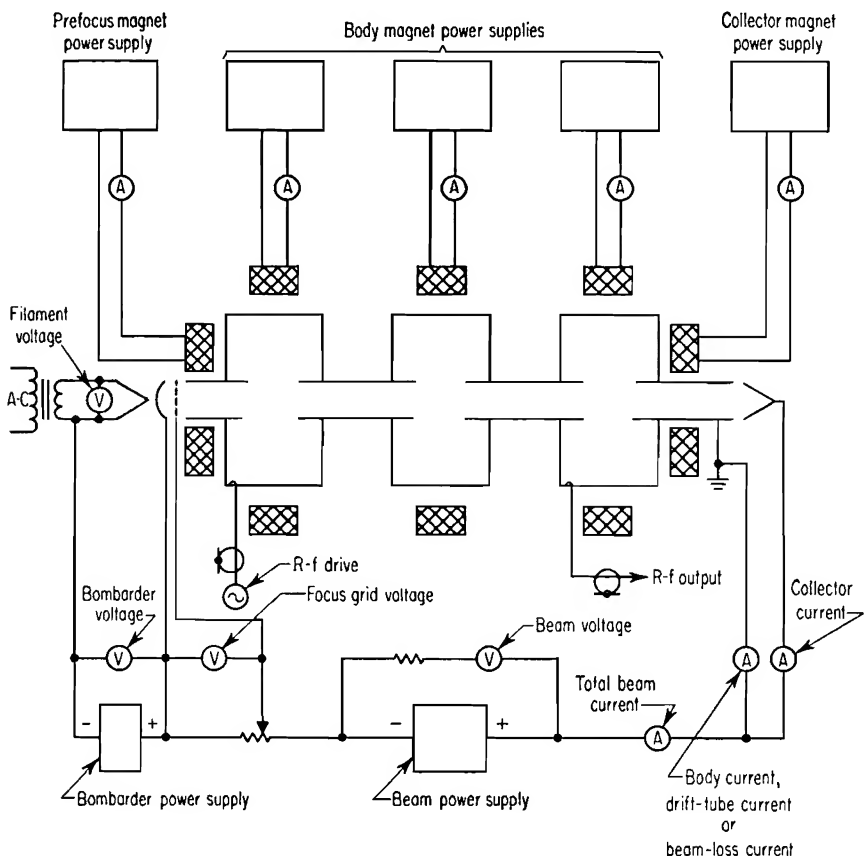


FIG. 4-4. Typical metering and power-supply connections for a klystron amplifier.

currents and hence the magnetic field are adjusted to minimize the number of electrons striking the walls of the drift tube. Electrons that do strike the walls of the drift tube at any point throughout the length of the klystron result in a current which is metered in a drift-current (sometimes called body current) meter shown in Fig. 4-4. This current does not normally exceed 5 per cent of the total beam current.

When a r-f signal is fed into the resonant input cavity, an r-f voltage will be set up across the gaps of the cavity. As the electron beam passes this gap, it will alternately be speeded up and slowed down according

to the polarity of the alternating r-f voltage across the gap. At this point, the beam is velocity-modulated. As the beam travels down the tube, the faster electrons will catch up to the slower electrons in the beam. At this point, the electrons will appear to be bunched together. When the beam is bunched in this manner, it is called *density modulation*. If another cavity and gap are placed at this point in the klystron, the bunching action will be reinforced, and the grouping of the electrons will be more discrete.

The energy can be extracted from the electron beam in much the same manner as the original velocity modulation of the electron beam. As the density-modulated beam passes the output cavity, the varying gap polarity will accelerate the less dense portions of the beam and slow up

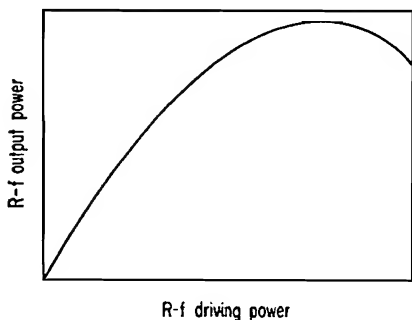


FIG. 4-5. R-f drive vs. output power for a klystron amplifier. (Courtesy of IRE.)

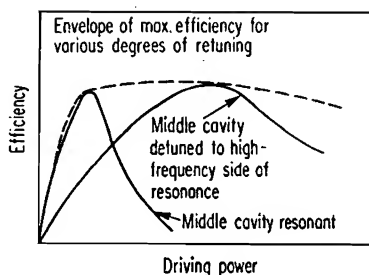


FIG. 4-6. Klystron amplifier efficiency vs. r-f drive depends on the tuning of the penultimate cavity.

the denser portions of the beam. This results in more electrons giving up energy than there are electrons taking energy. The r-f energy taken from the beam is transferred to the load circuit connected to the cavity.

Klystron Amplifier Characteristics. The klystron amplifier cannot be tuned in the manner typically associated with other amplifiers. The anode or collector current in the klystron does not vary with r-f drive, power output, or the tuning of the cavities. In this sense, it may be likened to a class A amplifier.

The klystron power output will vary with the r-f drive in the manner shown in Fig. 4-5. This curve is a mathematical characteristic of the klystron and is a Bessel function of the first order and first kind. The first portion of this curve is essentially linear, thus permitting the use of the klystron amplifier as a linear amplifier. In television applications, the r-f drive corresponding to the synchronizing pulse is usually stretched or amplified so that it will be correctly proportioned in the output.

The efficiency of the klystron amplifier depends to a large extent on the tuning of the penultimate cavity (the cavity preceding the output

cavity). This cavity is normally detuned to a frequency higher than the midband frequency. This characteristic is illustrated in Fig. 4-6. In klystrons with four or more cavities, the penultimate cavity should never be permitted to resonate at the operating frequency. This would cause the cavity to act like an unloaded output cavity. In turn, this would develop extremely high r-f voltages across the gap, which would result in arc-overs and cracking of the tube's ceramic walls.

In klystrons, it is also generally necessary to ensure that the output cavity is over-coupled to the load at all times; otherwise the high r-f voltages and currents developed may damage the tube.

The efficiency and power output of the klystron amplifier also depends on the collector voltage as shown in the curve of Fig. 4-7. The optimum bunching points depend on the speed of the electrons in the beam traveling down the drift tube. The points of optimum bunching also determine the original placement of the cavities along the drift tube by the manufacturer. If the collector voltage is changed, the speed of the

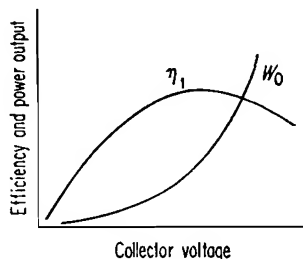


FIG. 4-7. Klystron amplifier efficiency and power output vs. collector voltage.

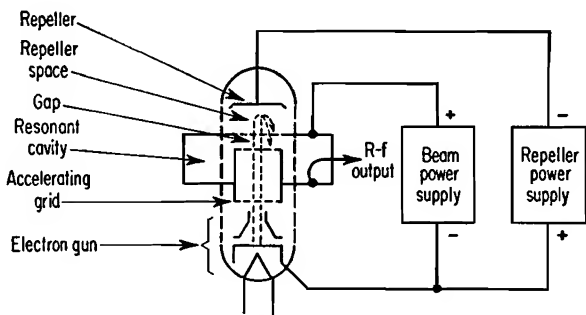


FIG. 4-8. Basic circuit of reflex klystron.

electrons in the beam will also change, and thus the location of the optimum bunching points in the drift tube will change. Thus there is one optimum collector voltage. The points of optimum bunching also depend on the frequency of operation. There is a broad fall-off on each side of the optimum frequency.

4-7. Reflex Klystrons. The reflex klystron is an oscillator and is widely used as the r-f power output in microwave relay transmitters as well as the local oscillator in receivers. The efficiency of the reflex klystron is in the order of 1 per cent. For this reason the reflex klystron is generally limited to applications requiring 10 watts of output power.

The reflex klystron uses the velocity-modulation principle described previously. Figure 4-8 illustrates the schematic of the reflex-klystron oscillator. The electron beam is attracted to the positively charged resonator cavity anode. The electron beam will pass through a gap or

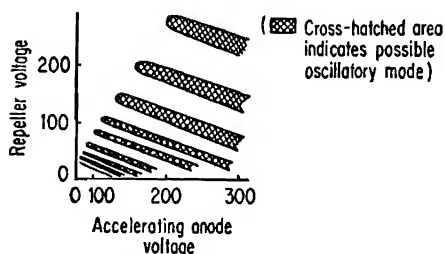


FIG. 4-9. Mode pattern for typical reflex klystron showing possible combination of repeller and anode voltage which will result in oscillation.

bunched electrons are slowed down by the alternating r-f resonator voltage which means that some energy is given to the gap, and so the r-f oscillation is sustained.

Oscillation in the reflex-klystron oscillator can only occur for certain anode voltages and repeller voltage combinations. Figure 4-9 illustrates the mode patterns for a typical reflex-klystron oscillator.

The r-f output frequency of the reflex klystron is primarily determined by the resonant frequency of the resonator. However, it will also depend on the velocity of the electrons and the repeller-space transit time. Thus the frequency is also somewhat dependent on the resonator anode voltage and the repeller voltage. A frequency variation in the order of 1 per cent is commonly achieved by intentionally varying the repeller voltage around a fixed value of anode voltage. Figure 4-10 illustrates the frequency and power output characteristics of a reflex klystron for changes in the repeller voltage.

grid in the resonator anode and will pass into the repeller space. The repeller element is negatively charged and will reverse the direction of the electrons back toward the resonator anode. However, as the electrons initially pass through the gap into the repeller space, they are velocity-modulated by the changing polarity of the resonator. This will cause the electrons to partially bunch at the r-f rate. The returning

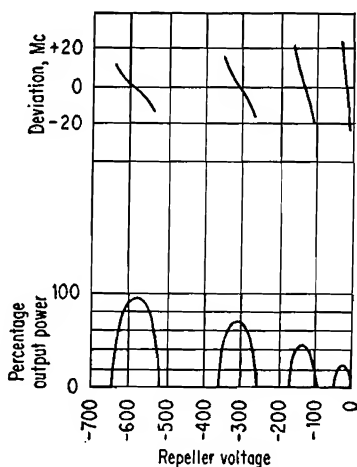


FIG. 4-10. Characteristics of typical reflex klystron. Power output and frequency vs. repeller voltage. (Courtesy of I.T.T.)

4-8. Traveling-wave Tubes. The traveling-wave tube is used as a broadband r-f amplifier at varying power levels up to several kilowatts

continuous wave. While special versions of the tube are in the form of an oscillator, it is most commonly used as an amplifier. In the traveling-wave tube, the electron beam is made to interact with the field of a wave traveling along a transmission line which is part of the tube.

The basic construction of a traveling-wave tube is shown in Fig. 4-11. Since the basic operation of the traveling-wave tube does not depend on the characteristics of any resonant circuits, the bandwidth of the tube is

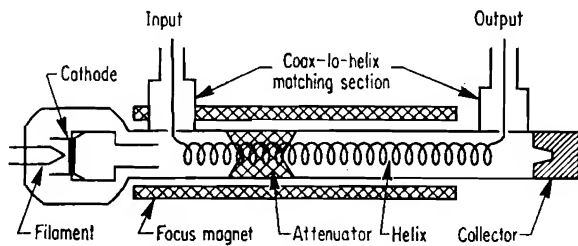


FIG. 4-11. Cross section of traveling-wave tube.

quite large. Bandwidths of an octave, i.e., 2:1, can be obtained at low power, while bandwidths of 10 per cent are common at the higher-power levels. The efficiency of the traveling-wave tube is usually in the order of 25 per cent, while power gains of 30 to 50 db are typical.

In operation, an electron gun emits electrons which are focused into a narrow beam. This beam is attracted toward the positively charged collector. The beam is focused or kept from spreading by an axial magnetic field supplied by an electromagnetic solenoid. This field is maintained over the entire length of the tube.

The r-f portion of the traveling-wave tube consists of a transmission line, typically a helically wound transmission line, wound over the beam as shown in Fig. 4-11. Electrically, the helix corresponds to many wave-lengths at the operating frequency of the amplifier. The phase velocity of the wave traveling along this line is one-tenth to one-third that of light. The velocity of the electron beam is adjusted to be approximately the same as the phase velocity of the wave traveling in the helix. This is achieved by adjusting the collector voltage. Figure 4-12 shows the relationship between the gain and collector voltage of a typical traveling-wave tube.

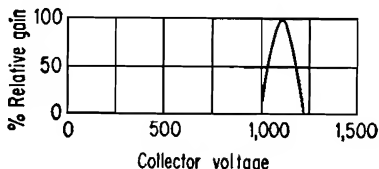


FIG. 4-12. Traveling-wave tube gain vs. collector voltage. (Courtesy of ITT.)

In practice, other slow-wave structures besides the helix structure previously described are possible. The only requirements are that the structure have a phase velocity that is a small fraction of the speed of

light and also that the structure must have an axial component of electric field. Figure 4-13 illustrates two other slow-wave structures that fulfill these requirements.

The traveling-wave tube will oscillate if some means of suppressing these oscillations is not incorporated into the tube's design. Any mismatch in the output circuit of the tube will cause a reflected signal to travel toward the input end of the tube and will result in continued oscillation. This tendency toward oscillation can be suppressed by incorporating an attenuator near the input end of the tube, as shown in

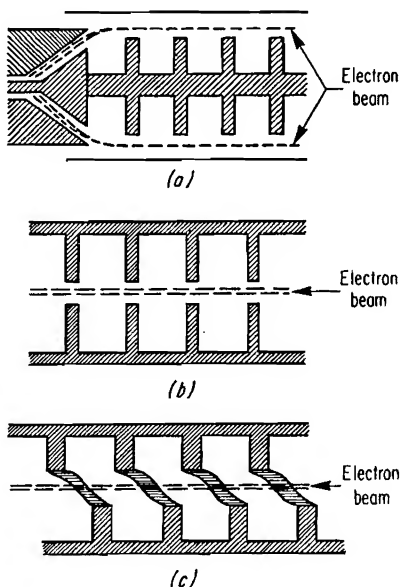


FIG. 4-13. Slow-wave structures used in traveling-wave tubes.

Fig. 4-11. The loss of the attenuator is generally made approximately equal to the forward gain of the tube. The attenuator, intended to absorb the reflected r-f power, will also absorb the forward or incident r-f power in the helix structure. However, since the attenuator is placed near the beginning of the tube, the power level is not ordinarily large. In addition, the electron-bunching process is not greatly affected by the attenuator since the beam induces the forward signal back into the helix after the attenuator.

The Carcinotron. The carcinotron is an oscillator which functions in much the same manner as the traveling-wave tube. The carcinotron, also known as a backward-wave oscillator, may use either the folded-line structure shown in basic form in Fig. 4-14, or it may use the helix structure previously described for the amplifier of the traveling-wave tube.

In this tube one end of the transmission-line structure is terminated in a nonreflecting load as shown in Fig. 4-14. If one assumes that the tube is in operation, a wave will be traveling from right to left in Fig. 4-14. The electric field produced by such a wave at the intersection of the electron beam from the gun and the transmission line will result in a bunching of the beam at this point. This process is much the same as the bunching process that takes place at the input gap of a klystron. As the partially bunched beam travels down the tube toward the collector, a similar action occurs each time the beam interacts with the wave in the folded transmission line. This will result in a continued oscillation.

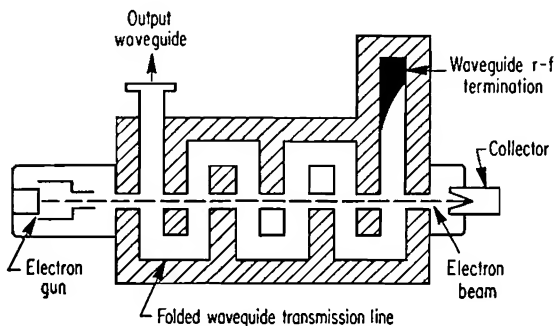


FIG. 4-14. Typical backward-wave oscillator (carcinotron).

The frequency of the oscillation in a carcinotron will depend on the voltage applied to the collector. Varying this voltage results in a frequency range of 2 to 1. Carcinotrons have been constructed in the frequency range of 300 Mc to 200 kMc. Power outputs of up to 150-kw pulse power and 1-kw continuous-wave (c-w) power have been developed. Efficiencies may be as high as 50 per cent, although some have efficiencies as low as 1 per cent.

4-9. Magnetrons. Magnetrons are microwave r-f oscillators commonly used in radar and electronic countermeasure applications. The r-f frequency range of the magnetron is from 500 Mc to 100 kMc. The efficiency of the magnetron is up to 70 per cent. The power outputs range up to 5 Mw in pulse applications and ten kw in c-w applications.

A typical magnetron is shown in the cutaway view of Fig. 4-15. The magnetron also requires a magnetic field parallel to the cathode surface. This field is usually supplied by a permanent magnet, although electromagnets are sometimes used.

With no axial magnetic field present in the magnetron cross section shown in Fig. 4-16 the electron path will be as shown in Fig. 4-16a, i.e., directly from the cathode to the anode. With increasing axial magnetic

field, the electron path will change from that shown in part (b) to that in part (c), and then to that shown in part (d). The electron path shown in Fig. 4-16c will not result in anode current since the electrons emitted by the cathode are all deflected back to the cathode. The magnetic field that produces this condition is called the *cutoff field*. When the

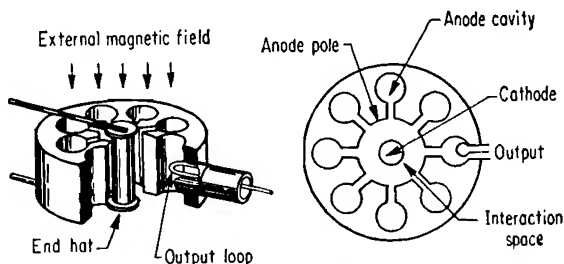


FIG. 4-15. Basic construction of a typical magnetron. (a) Cutaway view; (b) cross section perpendicular to axis of cathode.

field is adjusted beyond the cutoff magnetic field, the electrons interact with the resonant cavities. This sets up an r-f electric field across the gaps, which modifies the electron orbit. This results in a bunched space charge that effectively rotates around the axis of the tube. This excites the resonant cavities and results in continued oscillation. The

r-f output frequency of the magnetron will depend on the mode. There will be as many modes of operation as there are resonant cavities. The mode normally employed in the magnetron is called the π mode, which means that there is an r-f phase difference of 180° between adjacent poles. This mode of operation is assured by means of "strapping." Strapping is done simply by electrically connecting every other gap together as shown in Fig. 4-17. Mode suppression is also achieved by another anode structure known as the *rising sun*. This is shown in Fig. 4-18.

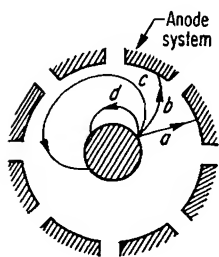


FIG. 4-16. Electron path in a magnetron. (From "Radio and Elect. Engineering," p. 692, McGraw-Hill Book Company, Inc., New York, 19.)

The performance of magnetrons is typically given in the form of two diagrams known as the *performance chart* and the *Rieke diagram*. These two diagrams, shown in Figs. 4-19 and 4-20, are essentially self-explanatory.

4-10. Platinotrons, Amplitrons, and Stabilotrons. The term *platinotron* is a general name for a crossed-field microwave tube utilizing the magnetron principle of interaction. A cross section of the platinotron is shown in Fig. 4-21. The platinotron can be used as either a high-power broadband amplifier or as a stable oscillator. As an amplifier, the platinotron

is sometimes called an *amplitron*; and when the platinotron is used as an oscillator, the tube is often called a *stabilotron*. The only difference in either case is the external circuitry connected to the tube.

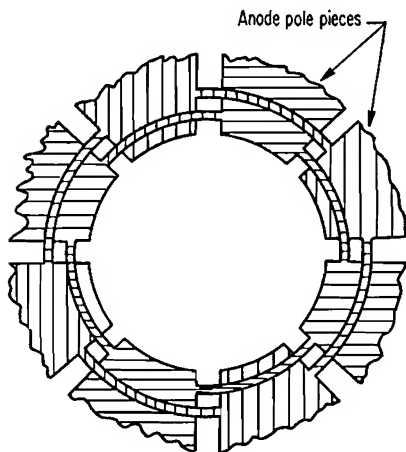


FIG. 4-17. Pole connections in a strapped magnetron.

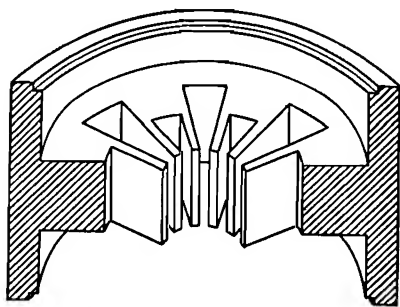


FIG. 4-18. Cross section of rising-sun magnetron anode.

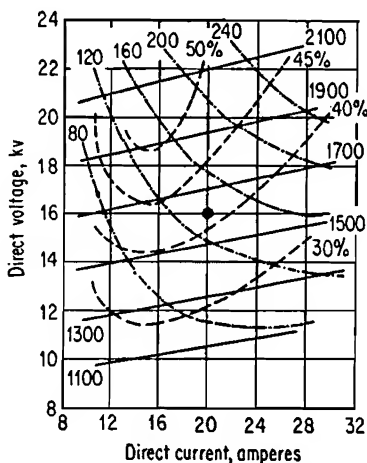
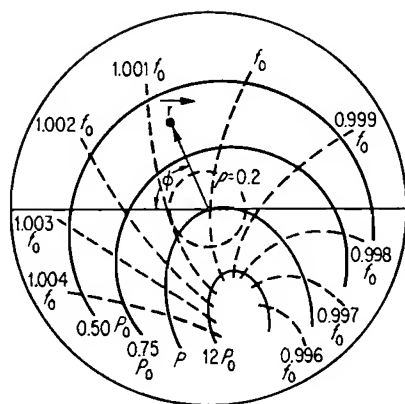


FIG. 4-19. Performance chart of a typical magnetron. (Courtesy of Bell System Tech. J.)

The platinotron is bidirectional, i.e., while the tube will exhibit a gain for a signal passing from input to output, it will also pass a signal from the output back to the input with little attenuation. Thus any reflected energy from the load must be minimized since it will appear at the output of the r-f driver of the platinotron.



———— Constant power output
 - - - - Constant frequency

FIG. 4-20. Rieke diagram of typical magnetron. (Courtesy of Bell System Tech. J.)

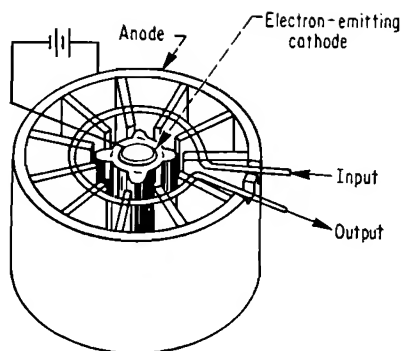


FIG. 4-21. Cross section of a typical platinotron. (Courtesy of Electronics.)

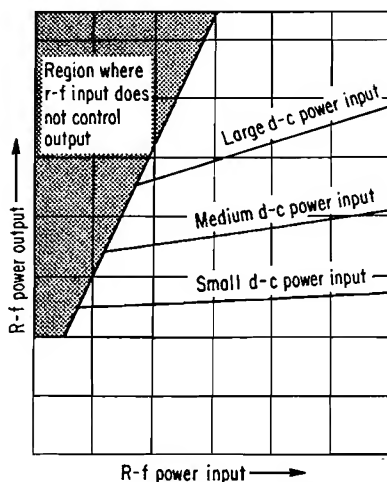


FIG. 4-22. Platinotron characteristics. (Courtesy of Electronics.)

The r-f input signal to a platinotron is not linearly related to the r-f output, and under some low-drive conditions the r-f input signal has no control on the power output. These characteristics are illustrated in Fig. 4-22. The gain of the platinotron varies from 5 to 20 db according to the r-f drive.

The platinotron is used in the frequency range of 200 Mc to 10 kMc. Platinotrons are capable of from 200 kw to 8 Mw of pulse power, and

up to 20 kw of c-w power. The efficiency of the platinotron is up to 70 per cent and may drop to 50 per cent for bandwidths of 10 per cent.

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COUPLING CIRCUITS

5-1. Coupling Networks. General. Coupling networks—typical of which are those found between an amplifier and transmission line, between a transmission line and antenna, or between stages in a transmitter—are required to perform a variety of functions. These functions are summarized as follows:

1. The coupling network must be capable of transforming the load impedance to the value required by the generator. The generator impedance in transmitter applications is usually either the impedance determined by the amplifier operating conditions (this was discussed in Chap. 3) or the impedance of a transmission line connected to the input of the coupling network.

2. The bandwidth of the network must be sufficiently wide for the application. At the same time, it is desirable that the coupling network suppress harmonics and other out-of-band or spurious signals that may appear in the transmitter output.

3. The efficiency of the coupling network should be as high as possible.

5-2. Parallel-tuned Circuits. At resonance, the parallel-tuned circuit exhibits a high impedance which makes it more convenient for use in the

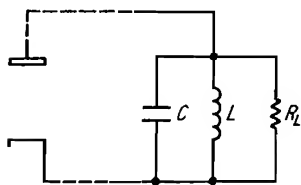


FIG. 5-1. Parallel-tuned circuit with load resistance R_L .

plate circuit of most tubes. The value of the impedance at resonance is essentially equal to the value of the load resistance R_L shown in Fig. 5-1. The load resistance seen by the amplifier is the transformed value of the actual load resistance. The actual load resistance for an amplifier may be 50 ohms, but it is transformed by an impedance-matching network to appear as the resistance value R_L

which is usually in the thousands of ohms. Ideally the value of R_L should equal the value calculated for the tube by the methods described in Chap. 3. The impedance of a parallel-tuned circuit at resonance is also approximately equal to a value of Q times the reactance of either the inductive or capacitive branches.

At frequencies below resonance, the parallel-tuned circuit acts like an

inductance; at frequencies above resonance, the circuit behaves like a capacitance. In cases where it may be necessary to calculate the impedance of a parallel-tuned circuit at frequencies other than resonance, it can be found by the equation

$$Z_p = \frac{R_L}{\sqrt{1 + R_L^2 \left(2\pi f C - \frac{1}{2\pi f L} \right)^2}} \quad (5-1)$$

where Z_p = magnitude of parallel-circuit impedance

R_L = parallel load resistance, ohms (shown in Fig. 5-1)

C = parallel capacitance, farads (shown in Fig. 5-1)

L = parallel inductance, henrys (shown in Fig. 5-1)

f = frequency of signal across LCR circuit of Fig. 5-1, cps

At resonance, all the coupling networks to be described can be reduced to a simple parallel-tuned circuit consisting of an inductance L , a capacitance C , and a load resistance R_L , as shown in Fig. 5-1.

The value of the inductance and the capacitance is determined by the desired operating Q of the tuned circuit. For all practical purposes, the operating Q of the amplifier may be defined as

$$Q = \frac{R_L}{X} \quad (5-2)$$

where R_L = effective load resistance in parallel with tuned circuit, ohms

X = reactance of either capacitor C or inductance L at resonant frequency, ohms

It is the usual practice in transmitter applications to make the operating Q of an amplifier somewhere between 10 and 20. Lower values result in increased harmonic output, while higher values result in increased losses as a result of the circulating currents in the tank circuit. Thus, with the value of Q fixed and the value of the load resistance known, the value of the capacitive and inductive reactance can be determined from Eq. (5-2). The actual value of the inductance and capacitance can be determined thus:

$$L = \frac{X_L}{2\pi f} \quad (5-3)$$

$$C = \frac{1}{2\pi f X_c} \quad (5-4)$$

where L = inductance, henrys

C = capacitance, farads

f = frequency, cps

X_L = reactance of inductance L , ohms

X_c = reactance of capacitance C , ohms

Circulating Current. In a parallel-resonant circuit, the normal interchange of current between the inductance and the capacitance will result in a circulating current which is Q times the input r-f line current to the parallel-tuned circuit. When the parallel-tuned circuit is in the plate circuit of a tube, the circulating current in the tuned-circuit components can be estimated from a knowledge of the Q and the plate current by the equations

$$I_{\text{cire}} \cong 2I_{\text{dc}}Q \quad (5-5)$$

where Q = value defined in Eq. (5-2)

I_{cire} = circulating current, amp

I_{dc} = d-c plate current, amp

Thus, the circulating current in a tank circuit may be in the order of amperes for even a low-power amplifier, and some care must be used in the selection of these components to ensure that excessive losses do not occur. For this reason, the Q of the individual components must be made as high as is practical.

In applications where the tuned circuit is not in the plate circuit of a tube, the circulating current can be approximated thus:

$$I_{\text{cire}} = \frac{E}{X} \quad (5-6)$$

where E = measured voltage across the parallel-tuned circuit

X = reactance of either capacitor or inductance in parallel-tuned circuit, ohms

I_{cire} = circulating current, amp

If the Q of the components in the tank circuit can be measured or is known, then the power lost in the tank circuit can be easily calculated. The power lost in a component (either a coil or capacitor) is

$$P = \frac{I_{\text{cire}}^2 X}{Q} \quad (5-7)$$

where I_{cire} = circulating current calculated from Eq. (5-4)

X = reactance of coil or capacitor at operating frequency

Q = figure of merit of coil or capacitor

P = power lost, watts

Bandwidth. The bandwidth of tuned circuits is usually expressed as the difference in frequency between the two points at which the power of the amplifier has dropped to one-half the midrange value. These points are commonly referred to as the *half-power points*. In voltage-response curves of the type usually observed on oscilloscopes, these half-power points are identified as the points where the voltage has dropped to 0.707 of the midrange value. This is shown in Fig. 5-2.

The bandwidth of an amplifier is determined by the load resistance across the tuned circuit and the output capacitance of the amplifier. Thus, for a single-tuned circuit, the bandwidth is

$$BW = \frac{1}{2\pi R_L C} \quad (5-8)$$

where BW = bandwidth between half-power points, cps

R_L = load resistance, ohms (appearing across tuned circuit shown in Fig. 5-1)

C = total capacitance of tuned circuit of Fig. 5-1, farads (This includes any stray capacitances, and output capacitances of the tube.)

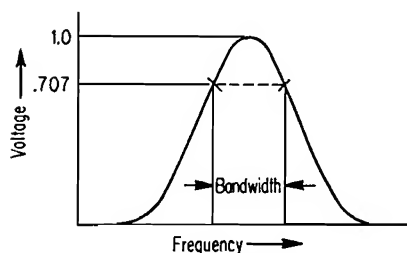


FIG. 5-2. The bandwidth as determined by the voltage response of a tuned circuit.

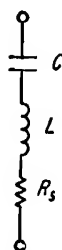


FIG. 5-3. Series-tuned circuit.

5-3. Series-tuned Circuits. The series-tuned circuit shown in Fig. 5-3 has the inverse characteristics of the parallel-tuned circuit. At resonance, the series-tuned circuit exhibits a low impedance, the value of which is equal to the total series resistance of the tuned circuit. Below the resonant frequency, the series-tuned circuit will act capacitive, while at frequencies above the resonant frequency, the circuit will act like an inductance. The magnitude of the impedance can be calculated from the more general equation

$$Z_s = \sqrt{R_s^2 + \left(2\pi fL - \frac{1}{2\pi fC}\right)^2} \quad (5-9)$$

where Z_s = magnitude of impedance at any frequency for a series-tuned circuit, ohms

R_s = series resistance, ohms (shown in Fig. 5-3)

L = series inductance, henrys (shown in Fig. 5-3)

C = series capacitance, farads (shown in Fig. 5-3)

f = frequency across series LCR circuit of Fig. 5-3, cps

The Q of a series circuit is defined as the following

$$Q = \frac{X_s}{R_s} \quad (5-10)$$

where X_s and R_s are, respectively, the reactance and resistance in ohms of the element shown in Fig. 5-3.

At resonance, the voltages across the series capacitor and inductance will be Q times the voltage applied across the series-tuned circuit. Thus,

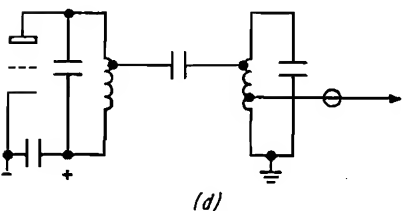
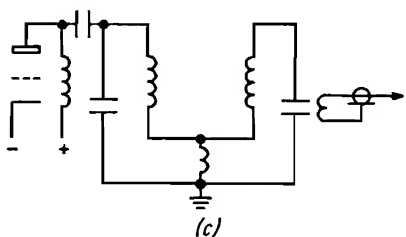
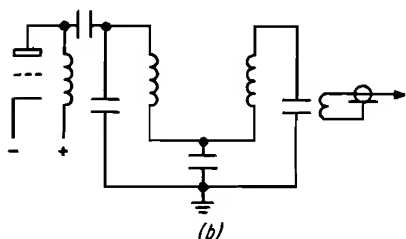
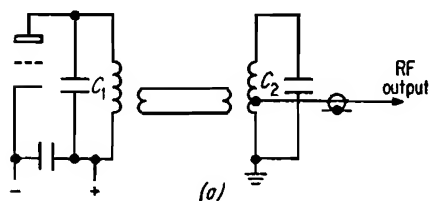


FIG. 5-4. Various methods of coupling between double-tuned circuits.

the mutual inductance provided in Fig. 5-4a, or through the common impedance shown in Fig. 5-4b and c, or through direct capacitance coupling as shown in Fig. 5-4d.

The bandwidth of the double-tuned circuit is greater by a factor of 0.707 for a circuit with the same input load resistance and with the same

if a circuit has a Q of 100, and only 1 volt is applied across the tuned circuit, the voltage across the inductor or the capacitor will be 100 volts. Thus in transmitting circuits, where the voltages are usually considerably higher than those in the example, some care must be exercised in choosing the proper components.

The bandwidth of a series-tuned circuit is conveniently expressed by the equation

$$BW = \frac{R_s}{2\pi L_s} \quad (5-11)$$

where R_s = series resistance, ohms
(shown in Fig. 5-3)

L_s = inductance of series-resonant circuit, henrys
(shown in Fig. 5-3)

BW = bandwidth between half-power points, cps

5-4. The Double-tuned Circuit.

In many applications, the frequency response of the single-tuned circuit is not adequate, and therefore the use of a double-tuned circuit is desirable. Figure 5-4 shows some of the double-tuned circuit configurations that may be found in transmitting equipment. The coupling between the primary and secondary may be either through

input capacitance as a single-tuned circuit. This can be noted by observing the following equation for the double-tuned circuit and comparing it with the bandwidth equation for a single-tuned circuit shown in Eq. (5-8).

$$BW = \frac{C_2}{2\sqrt{2}\pi R_L C_1} \quad (5-12)$$

where BW = bandwidth for double-tuned circuit, cps

R_L = load resistance, ohms (reflected across the primary of the double-tuned circuit shown in Fig. 5-4a)

C_2 = secondary output capacitance, farads (shown in Fig. 5-4)

C_1 = primary capacitance, farads (shown in Fig. 5-4a)

Equation (5-12) is based on the fact that the tube does not supply any appreciable loading on the primary circuit; i.e., the Q of the primary tuned circuit is essentially infinite. This is the usual case for pentodes and tetrodes where the internal plate resistance is very high.

In many cases, the amplifier is designed around a particular tube. This dictates the value of the load resistance R_L . If the amplifier is used in the vhf region, the tube construction will also dictate the output capacitance, and hence the value of C_1 . Thus for a given bandwidth, the only remaining unknown is C_2 . In these cases, Eq. (5-12) can be used to solve the value of C_2 .

The load resistance presented to the tube at midband by a double-tuned circuit can be found by

$$R_L = \frac{2R_2 C_2}{C_1} \quad (5-13)$$

where R_L = load resistance, ohm (reflected across the primary of the double-tuned circuit)

R_2 = secondary load resistance, ohms (effectively across the entire secondary shown in Fig. 5-4)

C_1 = primary capacitance, farads (as shown in Fig. 5-4)

Adjustment of Double-tuned Circuits. A double-tuned circuit contains four variables that require adjustment in order to secure the optimum performance from the circuit. These four variables are as follows:

1. The primary resonant frequency
2. The secondary resonant frequency
3. The primary-secondary coupling
4. The secondary loading resistance

These adjustments may be performed most conveniently by displaying the amplitude response of the tuned circuit on an oscilloscope. In this way, the results of any changes in the tuned-circuit adjustment can be immediately observed.

Normally, the primary and secondary resonant circuits are tuned to the same frequency. These circuits can be initially set to the same frequency by means of a grid-dip meter. However, it is also necessary to disable the coupling between the primary and secondary circuits or actually short-circuit the tuned circuit not being adjusted.

With the secondary circuit lightly loaded (for example, the primary tap of Fig. 5-4*d* close to ground) the coupling between the primary and

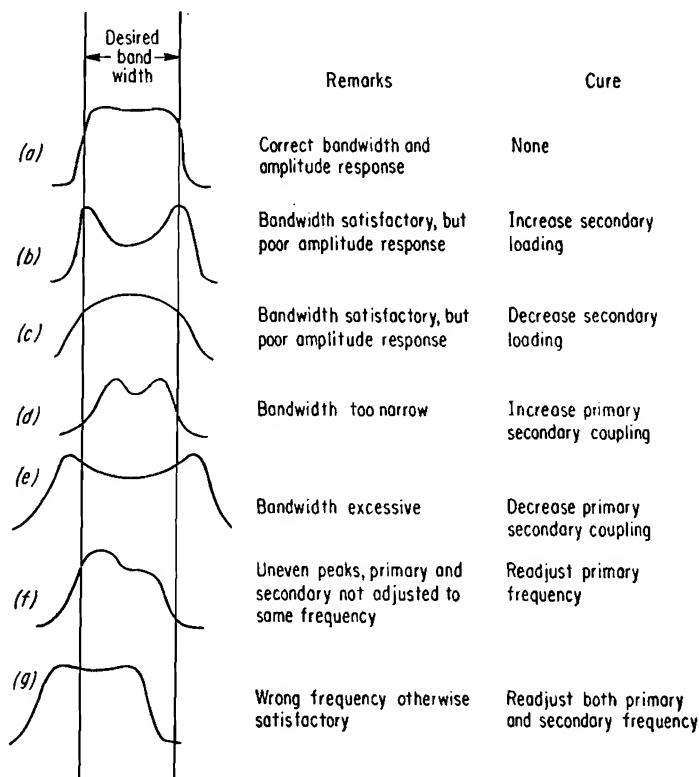


FIG. 5-5. Amplitude response curves of double-tuned circuits.

secondary is adjusted until a double-peaked response similar to that shown in Fig. 5-5 is obtained. If the peaks are not identical in amplitude, then the tuned circuits are not tuned to the same frequency. The higher peak is normally the primary, since it is the higher Q of the two. Adjust the resonant frequency of the primary and secondary until the response is centered in the desired band. The primary-secondary coupling is adjusted until the peaks are separated by approximately the bandwidth desired. The secondary loading is then increased until the response is filled in between the peaks. Some additional adjustment of all four

variables will be required to achieve the desired response. Figure 5-5 gives the typical response of some correctly and incorrectly adjusted double-tuned circuits, together with their defects and suggested cures.

5-5. Envelope Delay. Envelope delay is a complex principle which is not easily visualized. However, it is an important practical consideration in many tuned-circuit applications. This is particularly true in television applications where excessive variations in envelope delay that are uncompensated will result in overshoot and ringing in the final picture.

Basically, envelope delay is an indication of the transmission time of the signal through the system. More specifically, *envelope delay* can be defined as the measurement of the slope of the phase-vs.-frequency curve. This is mathematically expressed

$$T_e = \frac{\theta_2 - \theta_1}{2\pi(f_2 - f_1)} \quad \text{sec} \quad (5-14)$$

where θ_1 = phase shift at frequency f_1

θ_2 = phase shift at frequency f_2

f_1 = any sampling frequency at which the envelope delay is to be measured, cps

f_2 = sampling frequency very close to f_1

Envelope delay in tuned circuits would be a constant number if the phase shift through the tuned circuit at all frequencies were linearly related to the frequency. This would be the case if a plot of phase vs. frequency through the network resulted in a straight line. However, a plot of phase vs. frequency in most networks results in something quite remote from a straight line. A change in the slope of the line plotting phase vs. frequency also means that the envelope delay has also changed.

In a TV or other a-m signal, a shift in the phase of the sideband frequencies will also result in an equal shift in the phase of the r-f envelope of the signal. When the envelope is detected, the resultant demodulated signal also possesses this same amount of phase shift. If all the signals were delayed by the same amount of time, there would be no net change in the signal. However, if the transmission time (because of a nonlinear phase-shift-vs.-frequency characteristic of the tuned circuit) is different for different frequency sidebands, then time-delay distortion will result.

The envelope delay of a single-tuned circuit and some double-tuned circuits with varying degrees of coupling is shown in Fig. 5-6. The envelope delay of the double-tuned circuits depends on the ratio of the actual coupling between the primary and secondary of the tuned circuits to the critical coupling values.

5-6. Matching Network Configurations. Many types of networks are used to transform one impedance to another. However, in transmitter

applications, these networks are generally of the type known as L, π , or T networks, or are of the transformer or link-coupled types. Each type has its particular use and application, and each will be discussed in detail in the following material.

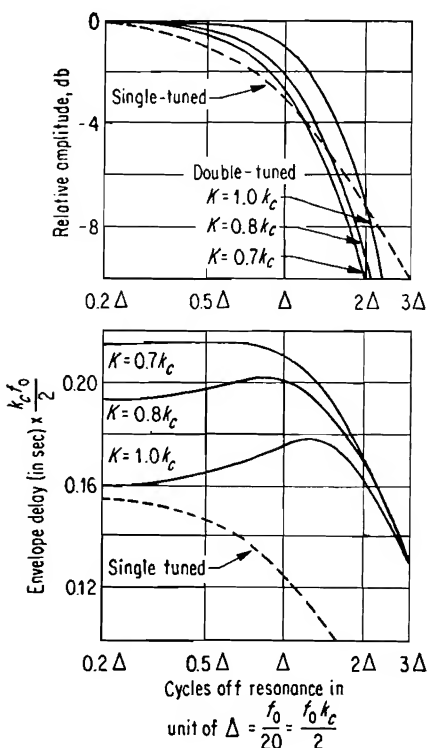


FIG. 5-6. Amplitude-response and envelope-delay curves for single- and double-tuned circuits.

The magnitude of the impedance $|Z|$ across the terminals of the series circuit shown in Fig. 5-7a is

$$Z = \sqrt{R_s^2 + X_s^2} \quad (5-15)$$

where R_s and X_s are in ohms and are shown in Fig. 5-7a. The operating Q of such a series circuit is

$$Q = \frac{X_s}{R_s} \quad (5-16)$$

Similarly, the magnitude of the impedance $|Z|$ across the terminals of the parallel circuit shown in Fig. 5-7b is

$$Z = \frac{R_p X_p}{\sqrt{R_p^2 + X_p^2}} \quad (5-17)$$

5-7. The Basis of the L, π , and T Matching Networks. *Series-to-parallel Circuit Equivalents.* The reason these networks are able to transform one impedance to another is that, for any series circuit consisting of a series reactance and resistance, there can be found an equivalent parallel network which possesses the same impedance characteristics. This is shown in the following equations.

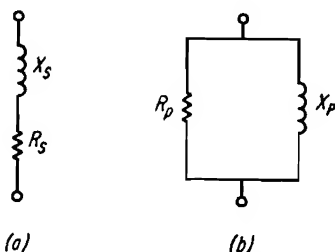


FIG. 5-7. Equivalent series and parallel LR networks described in Sec. 5-7. (a) Series; (b) parallel.

where R_p and X_p are in ohms and are shown in Fig. 5-7b. The operating Q of such a parallel circuit is

$$Q = \frac{R_p}{X_p} \quad (5-18)$$

Equations (5-15) and (5-17) can be equated, and Eqs. (5-16) and (5-18) can be substituted to produce the result

$$\frac{R_p}{R_s} = Q^2 + 1 \quad (5-19)$$

This equation illustrates the fact that any two resistances can be matched simply by adjusting the Q of the series or parallel reactance. For example, if it is desired to transform a 50-ohm resistance to 2,500 ohms, according to Eq. (5-19) the Q of the circuit should be 7. From Eq. (5-16) the series reactance to achieve this transformation should be

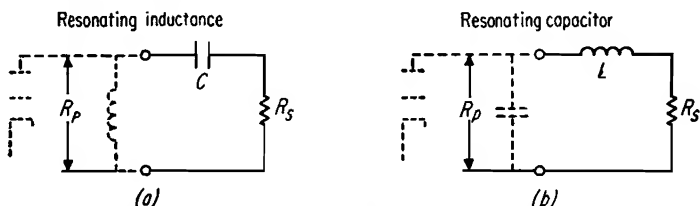


FIG. 5-8. Two equivalent L matching networks described in Sec. 5-8.

350 ohms. Thus, the series circuit consisting of a 50-ohm resistor and a 350-ohm reactance is equivalent to a parallel circuit consisting of a 2,500-ohm resistor and 357-ohm reactance [from Eq. (5-18)].

The reactance can be either a capacitor or inductance. The transformation is accomplished in either case.

5-8. The L Network. The L network is the first extension of the series-parallel impedance transformation described in Sec. 5-7 for use in practical transmitting equipment. It is the simplest of the matching networks and may take either of the two forms shown in Fig. 5-8. Either of these two networks will match equally well for any condition within the capability of the L-network configuration. The choice of which circuit to use is made on the basis of other component and circuit considerations apart from the impedance-matching characteristics. However the circuit shown in Fig. 5-8b is generally preferred because of its harmonic suppression characteristic. The shunt-resonating capacitor provides a bypassing path for the higher harmonic frequencies.

L-network Calculations. The steps required to successfully calculate the values required for any impedance-matching L network is as follows:

1. With the two impedances to be matched by the L network known, the required value of Q can be calculated by using Eq. (5-19).

2. The value of the series reactance can then be calculated by substituting the value of Q determined in step 1 in Eq. (5-16).

3. The value of the shunt reactance, which must be of opposite sign from step 2, can then be calculated by substituting the value of Q determined in step 1 in Eq. (5-18).

4. For either of the circuits shown in Fig. 5-8, the value of the inductance and capacitance to be used can be determined from the equation

$$L = \frac{X_L}{2\pi f} \quad (5-20)$$

$$C = \frac{1}{2\pi f X_c} \quad (5-21)$$

where L = inductance, henrys

C = capacitance, farads

X_L = reactance, ohms, calculated for the inductance

X_c = reactance, ohms, calculated for the capacitance

f = frequency, cps

Limitations of the L Network. One of the most serious limitations of the impedance-matching networks is that the operating Q of the circuit is determined solely by the ratio of the two impedances to be matched.

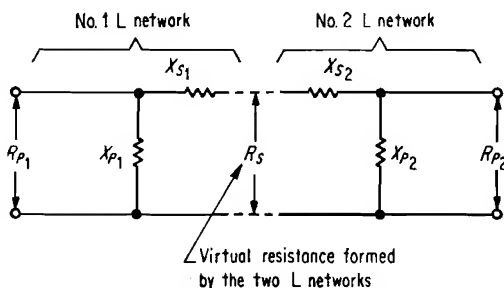


FIG. 5-9. A π network is made up of two back-to-back L networks.

In most practical amplifiers, it is desirable to hold the output circuit Q to around 10 to 20. Low Q 's result in increased harmonic output, while higher Q 's result in higher circulating currents and increased losses in the tank circuit. Therefore, if the ratio of the impedances to be matched result in an operating Q much beyond the limits of 10 to 20, it is desirable to use another type of matching network.

5-9. The π Network. The π network, so named because of its circuit resemblance to the mathematical symbol π , can be considered as two back-to-back L networks as shown in Fig. 5-9. Therefore much of the mathematics described in Sec. 5-8 on the L network is equally applicable to the design of the π network. The type of L network used is only limited by the requirement that the shunt and series arms be made up of

opposite types of reactance, i.e., if the shunt arm is capacitive, then the series arm must be inductive, and vice versa. With this in mind, the π networks shown in Fig. 5-10 can be worked out. Each of these networks is equally capable of performing the matching function within the capabilities of the π -network design limitations. However, as with the L network, the deciding factor for using any of these networks is often determined by the other circuit and component considerations. However, it is always true that networks employing shunt-capacitive arms

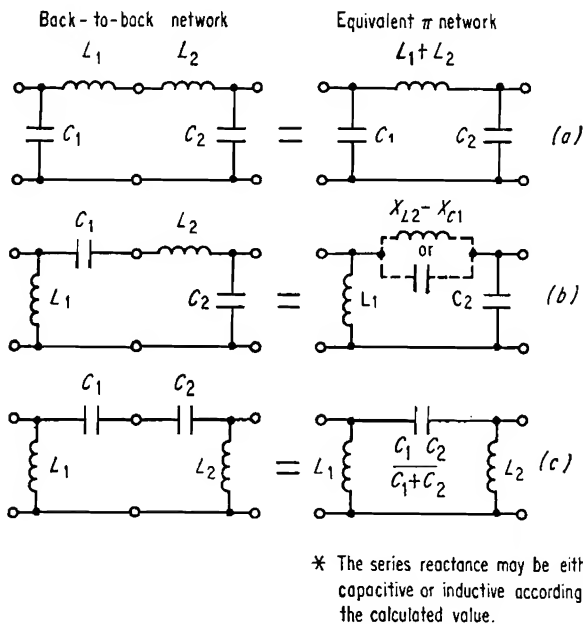


FIG. 5-10. Back-to-back L networks and equivalent π networks.

are preferred because of their inherent harmonic frequency-bypassing action.

π -network Calculations. The steps in designing a π network, based on a previously determined value of Q and the two impedances to be matched, is as follows:

1. Determine the value of input shunt reactance X_p according to Eq. (5-18).

2. From Eq. (5-19) determine the "virtual" resistance value. The virtual reactance is R_s in this equation. This value must be smaller than either of the two impedances to be matched. If it is not, then the value of Q previously selected must increase.

3. From Eq. (5-16) determine the value of the series reactance X_s .

4. The value of Q for the second or output L section can be determined by considering the virtual resistance, calculated in step 2 as R_v in Eq. (5-19).

5. Determine the series reactance X_s of the output L network from Eq. (5-16).

6. Determine the shunt reactance X_p of the output L network from Eq. (5-18).

7. The two series reactances of the L networks can be added together to form the complete π network.

8. Calculate the capacitance and inductance values that form the π network from the Eqs. (5-20) and (5-21).

5-10. The T Network. The T network can be considered as two back-to-back L networks arranged in much the same manner as the π network previously considered. The exception in this case is that the T network

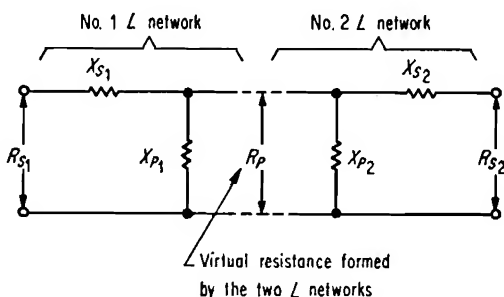


FIG. 5-11. A T network is made up of two back-to-back L networks.

has the shunt reactances of each hypothetical L network connected in parallel as shown in Fig. 5-11. Figure 5-12 also illustrates the various forms that an impedance-matching T network can assume. As pointed out for both the L and π networks, the form that the network takes has no relationship to its impedance-matching properties. All the networks shown in Fig. 5-12 will match with equal capability. The choice is simply in the practical choice of components, harmonic suppression characteristics, the desire or need for a d-c feed path, or a path to ground, etc.

T-network Calculations. The calculations follow the same general pattern of the π network, except that the virtual resistance formed at the junction of the two L networks must be higher than either of the two resistances being matched.

The calculations of the following steps assume that the value of Q and the two values of resistances to be matched are known.

1. With the value of Q and the input resistance to the first L network known, the virtual resistance can be calculated from Eq. (5-19). The virtual resistance will be R_v in this equation.

2. The value of the shunt reactance X_p of the first L network can be determined from Eq. (5-18).
3. The value of the series reactance X_s of the first L network can then be determined from Eq. (5-16).
4. The Q required for the second L section is determined from Eq. (5-19).
5. The value of the shunt reactance X_p of the second L network can be determined from Eq. (5-18).

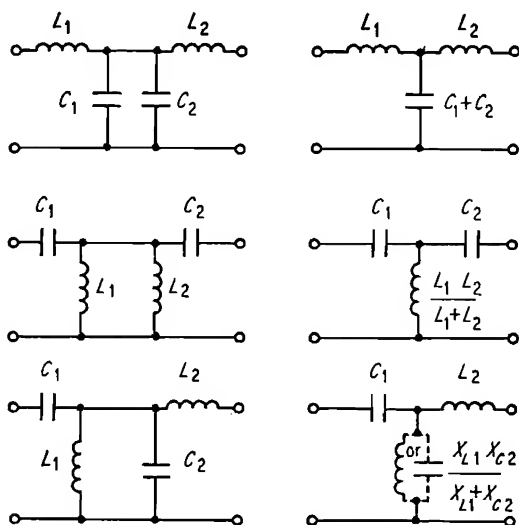


FIG. 5-12. Back-to-back L network and equivalent T network.

6. The value of the series reactance X_s of the second L network can be determined from Eq. (5-16).
7. The parallel reactances of the two L networks are combined to form one reactance X_{pt} by the equation

$$X_{pt} = \frac{X_{p1}X_{p2}}{X_{p1} + X_{p2}} \quad (5-22)$$

where X_{p1} = value of shunt reactance of first L network, ohms

X_{p2} = value of shunt reactance of second L network, ohms

8. The capacitance and inductance values are then calculated from Eqs. (5-20) and (5-21).

5-11. The π -L Network. In some cases, it may not be possible to match the specified input and output impedances at the desired input Q value with either the L, π , or T networks described in previous sections. In these cases, if it has been determined that none of the fixed parameters (input and output impedance and input Q) can be changed, then it is

necessary to use an additional L section to achieve the desired transformation. One of the more popular network configurations for achieving this is the π -L network. This is made up of a π network followed by another L section as shown in Fig. 5-13. Calculation of the network values for this configuration follows the principles previously outlined in Secs. 5-8, 5-9, and 5-10 and need not be repeated here. This network, in addition to the wider impedance-matching range also offers additional selectivity for the amplifier stage.

5-12. Reactive Load Impedances. In the previous examples of calculating L, π , and T networks, the loads were all assumed to be purely resistive. Very often this is not the case; that is, the load also contains some reactive components in addition to the resistance. In order to properly design a network of this type, it then becomes necessary to

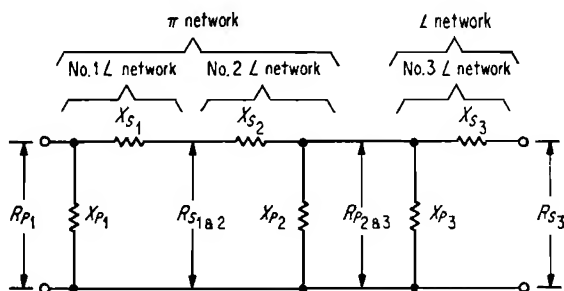


FIG. 5-13. A π L network made up of a combination of L networks.

determine what the reactive components of the load are, both in magnitude and types, i.e., inductive or capacitive. These reactive components can then be considered as part of the matching network. For example, if a π network were used to match into a tube, the tube input capacity as well as the stray capacity would be subtracted from the value calculated for the output network capacity. This would be the physical value required in the output of the network. If the matching network is to supply a mismatched transmission line, then the impedance existing at the end of the transmission line (the output of the matching network) can be most easily determined through the use of a Smith chart. In many cases, the degree of mismatch is unknown, and the worst condition must be assumed. The maximum value of the reactance for a mismatched 50- and 75-ohm transmission line is illustrated in the chart of Fig. 5-14.

5-13. Link-coupled Circuits. Link coupling is a commonly used method of matching impedances in transmitter applications and is particularly adaptable to applications requiring a match between a relatively high impedance to a low impedance. A typical application of this type

is between the output tank circuit of an r-f amplifier and a low-impedance transmission line. Some typical link-coupled circuits are shown in Fig. 5-15.

Tuned Link-coupled Circuits. The tuned link circuit affords some practical advantages over the untuned link circuit. Since the link reactance is canceled or tuned out, a given link coil can be satisfactorily used over a wider frequency range than an untuned link. In addition, the variable capacitor in series with the link provides a convenient method of adjusting the coupling without repositioning the link.

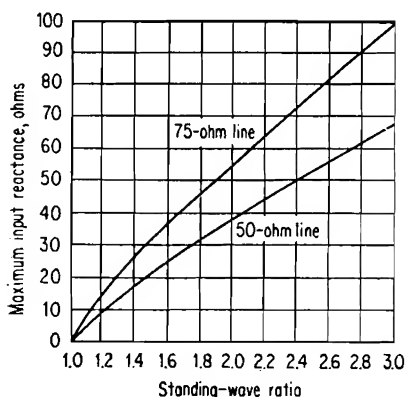


FIG. 5-14. Maximum input reactance for various standing-wave ratios for a 50- and 75-ohm transmission line. (Courtesy of ARRL.)

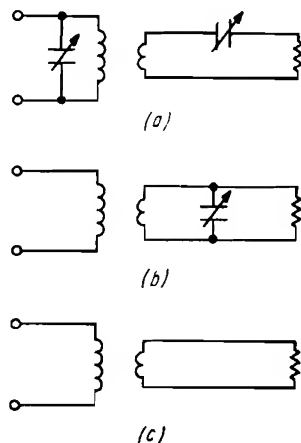


FIG. 5-15. Some typical link-coupled circuits. (a) Series-tuned link; (b) parallel-tuned link; (c) untuned link.

The degree of coupling which will present the required load resistance to the tube can be calculated by the equation

$$K = \frac{1}{\sqrt{Q_p Q_s}} \quad (5-23)$$

where Q_p = desired operating Q of primary tuned circuit shown in Fig. 5-15. This value is calculated from Eq. (5-2).

Q_s = operating Q of secondary tuned link circuit shown in Fig. 5-15. This value is calculated from Eq. (5-2) for the parallel-tuned link circuit or from Eq. (5-10) for the series-tuned link circuit.

K = coefficient of coupling expressed as a number less than 1.

In practical link-coupled r-f power circuits, the coefficient of coupling usually runs no more than 0.65. Some substitution of practical numbers in Eq. (5-23) reveals that, for a typical primary tank circuit Q of 10 to 20,

and for a typical maximum coefficient of coupling of 0.4 to 0.65, the Q of the secondary circuit may run between 0.12 and 0.63.

In practice, substitution of secondary Q 's in this region in Eq. (5-10) may result in excessively large values of series capacity. In these cases, the value can be reduced to a practical value, and the link inductance increased to maintain a resonant condition. This will result in an increase in the Q of the secondary, with a consequent reduction in the coefficient of coupling required. This latter merely means that the secondary link must be decoupled from the primary.

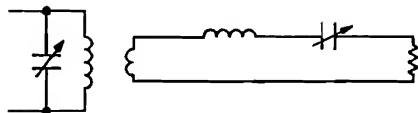


FIG. 5-16. The use of inductance in series with the link to obtain resonance.

In many cases, the increased inductance required is not added to the link but merely inserted in series with the link to bring the series circuit into resonance. This is shown in Fig. 5-16. In this case, the coefficient of coupling remains the same, although the Q is still increased as before.

The choice of using a parallel-tuned link or a series-tuned link is usually determined by other practical considerations apart from impedance matching. For example, where a d-c path is desirable (as for lightning protection) then the parallel-tuned type is desirable. In some circuits, because of the tuned-circuit construction, it may be necessary to provide for increased coupling to the amplifier tank circuit. In these cases the series circuit may be used, since for a given Q more link turns are required than for the equivalent parallel-tuned link circuit.

The parallel circuit will also have a lower r-f voltage across the secondary capacitor. This permits the use of lower voltage-rated components in the output circuit of the parallel-tuned link circuits.

Untuned Link-coupled Circuits. The untuned link circuit is shown in Fig. 5-15. For optimum operation, the link coil should have a reactance equal to the impedance of the transmission line or load to which it is coupled at the frequency at which the amplifier is operating. The optimum inductance of the link coil is given by

$$L = \frac{Z}{2\pi f} \quad (5-24)$$

where Z = transmission-line impedance, ohms

f = frequency of operation, cps

L = link inductance, henrys

Untuned link coils must be capable of close coupling to the amplifier tank circuit or primary, particularly if the standing-wave ratio of the transmission line to which the link is connected is low. Varying the coupling, i.e., the position of the untuned link, will require that the

primary or tank circuit be slightly retuned to compensate for the change in leakage reactance between the tank coil or primary and the link coil or secondary.

If any shunt capacitance becomes part of the link circuit, its reactance must also be considered. The net reactance of the shunt capacitance and the link reactance should still equal the load resistance.

In these cases, the link coil reactance can be found from the equation

$$X_L = \frac{ZX_c}{Z + X_c} \quad (5-25)$$

where Z = transmission-line impedance, ohms

X_c = shunt reactance, ohms

X_L = link reactance, ohms

The value of the link inductance can then be found by substituting the reactance found from Eq. (5-25) into Eq. (5-3).

The shunt-capacitive reactance must be considered in an untuned link circuit when the link is constructed from a coaxial transmission line in the general manner shown in Fig. 5-17. Links of this type are often used above 14 Mc to reduce the capacitive coupling between the amplifier tank coil and the link. A coaxial or shielded

link of this type reduces the radiation of any harmonic frequencies present in the amplifier output circuit along the outer conductor of the transmission line.

5-14. Balanced-to-unbalanced Coupling Networks. The previous discussion of coupling networks was concerned with the coupling of circuits in which both the inputs and outputs were unbalanced; i.e., one side of both the input and output circuits were common or grounded. However, it is sometimes necessary to couple between two circuits one of which is balanced and the other unbalanced. A balanced-to-unbalanced coupling network is required to couple between a single-ended amplifier stage and a push-pull stage (or vice versa), or between the output of a single-ended amplifier stage and a balanced transmission line.

The simplest method of coupling between an unbalanced circuit and a balanced circuit is by means of any of the link-coupled circuits previously described in Sec. 5-13. In this type of circuit, the load can be of either the balanced or the unbalanced type. However, in the L, π , or T networks previously described, the problem is somewhat more complicated

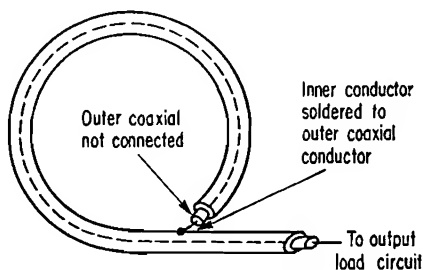


FIG. 5-17. The construction of an untuned coaxial link.

since there is no transformer coupling present as in the link circuit to provide the balanced output. In these types of networks, where a balanced output is desired, it is necessary to follow the L, π , or T network with an additional coupling network to provide the balanced output conditions. The balanced-to-unbalanced networks used to achieve this are commonly called *baluns*.

The development of a balun is best illustrated by the series of diagrams shown in Fig. 5-18. Fig. 5-18a illustrates a basic unbalanced circuit consisting of a generator and a load resistor. Some hypothetical voltage values as well as the polarities of these voltages are shown. An equivalent balanced system is shown in Fig. 5-18b. Note that the voltage of each generator is one-half the value shown in Fig. 5-18a and that each load resistance is also one-half the value shown in Fig. 5-18a. One of the generators shown in the balanced system of Fig. 5-18b can be replaced by connecting the remaining generator through a 180° phase-shifting network as shown in Fig. 5-18c. Thus an unbalanced generator is supply-

ing a balanced load by virtue of the 180° phase-shifting network. This, then, is the basic purpose of the balun network.

It is the characteristic of all baluns utilizing the principles described above that there is an impedance transformation ratio of 4:1 between the balanced and unbalanced terminals. This is assuming that there is no impedance transformation taking place in the phase-shifting network. This can be observed from the hypothetical voltages and resistances assumed in Fig. 5-18. The load resistance seen by the generator in Fig. 5-18a is four times the load resistance seen by the generator in Fig. 5-18c. Thus, to maintain the same load resistance on the generator in both cases, the output load on the balun must be four times the unbalanced load on the generator.

The 180° phase shift required to obtain the balanced load condition may be achieved by a variety of combinations of the L, π , and T networks or by the use of transmission-line sections with similar characteristics.

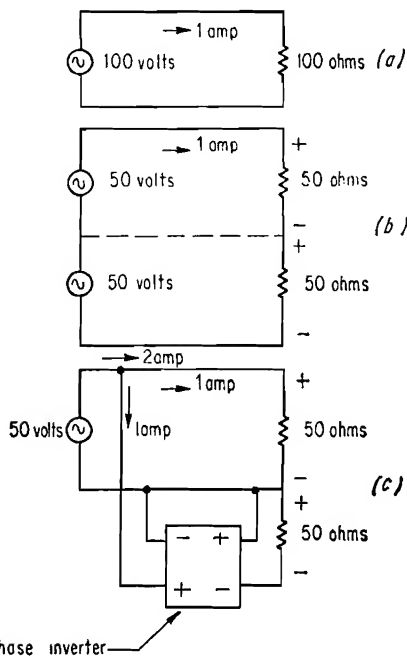


FIG. 5-18. The development of a balanced-to-unbalanced coupling network.

Figure 5-19 illustrates a practical network of this type using a π -L network. Note that one-half the balanced load resistor is effectively across the input to the balun network. The π -L network is used to shift the phase of the voltage of the remaining half of the balanced load by 180° .

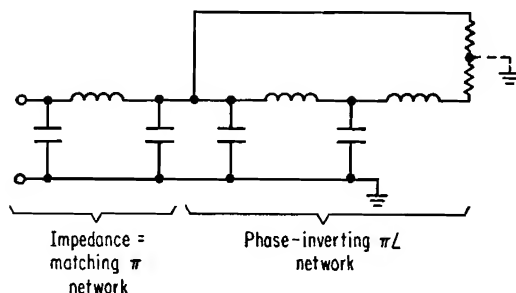


FIG. 5-19. A balanced-to-unbalanced coupling network made up from a π -L network.

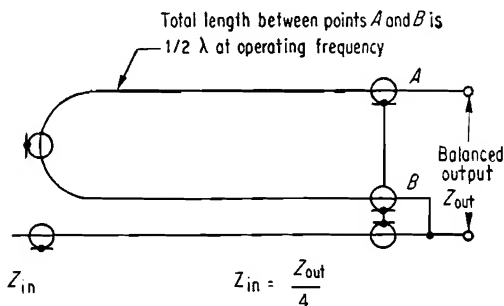


FIG. 5-20. A transmission-line balun.

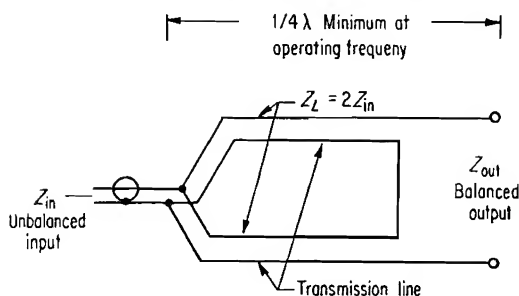


FIG. 5-21. A transmission-line balun essentially independent of frequency.

The equivalent balun made up of some transmission line is shown in Fig. 5-20. In this balun, the 180° phase shift is achieved by means of a one-half-wavelength section of transmission line. Note the inherent 1:4 impedance transformation of this balun.

The transmission-line balun of Fig. 5-21 achieves the 180° phase shift with the added advantage that it is essentially independent of the fre-

quency of operation. The only requirement is that the length of the transmission line making up the balun be long enough to act like a transmission line at the operating frequency. As a rule of thumb, it is usual to make the length of the balun sections not shorter than one-quarter wavelength at the frequency of operation. The length may be made as long as desired, providing the losses in the transmission line do not become excessive.

5-15. Matching-network Efficiency. The efficiency of the matching network is determined by the components used in the network. In general, the efficiency of any network is closely related to the Q of the components and in particular to the Q of the coils. The losses in the associated capacitors are usually negligible when compared to those of the coils.

The general statement of the efficiency of any matching network is

$$\% \text{ efficiency} = \frac{P_{\text{out}}}{P_{\text{in}}} \times 100 = \frac{P_{\text{out}}}{P_{\text{out}} + P_{\text{lost}}} \times 100 \quad (5-26)$$

where P_{in} = power into input of network, watts

P_{out} = power delivered from network to load, watts

P_{lost} = power lost or dissipated in network components, watts

A more specific equation of efficiency will depend on the network configuration. For an L network, the efficiency can be stated as

$$\% \text{ efficiency} = \frac{R_L}{R_L + R_s} \quad (5-27)$$

where: R_L = load resistance for L network, ohms

R_s = series loss resistance of matching network, ohms [determined from Eq. (5-10)]

Since the L network is the basic building block of the π and T networks, the product of the efficiencies of the individual L networks that make up the more complex network will give the over-all efficiency of that particular network.

From Eq. (5-27) it can also be observed that it becomes increasingly difficult to achieve high efficiencies for low values of load resistance. The load resistance tends to become comparable to the coil r-f resistance in these cases. This is the problem commonly encountered when attempting to match the output of an amplifier to antennas appreciably shorter than a quarter wavelength. The efficiency of an impedance-matching network for such an application is expressed by the

$$\% \text{ efficiency} = \frac{Q_L}{Q_A + Q_L} \quad (5-28)$$

where Q_L = ratio of coil reactance to coil resistance

Q_A = ratio of antenna reactance to antenna resistance

If one considers that a one-eighth-wavelength antenna usually has a Q of over 100, a network with coils with a Q of 100 would result in an efficiency of only 50 per cent; i.e., one-half of the transmitter output power would be dissipated in heating up the matching coil.

The efficiency of untuned and tuned link circuits can be determined by first determining the individual power losses in each of the coils. The sum of all the power losses can then be substituted in Eq. (5-26) to determine the network efficiency.

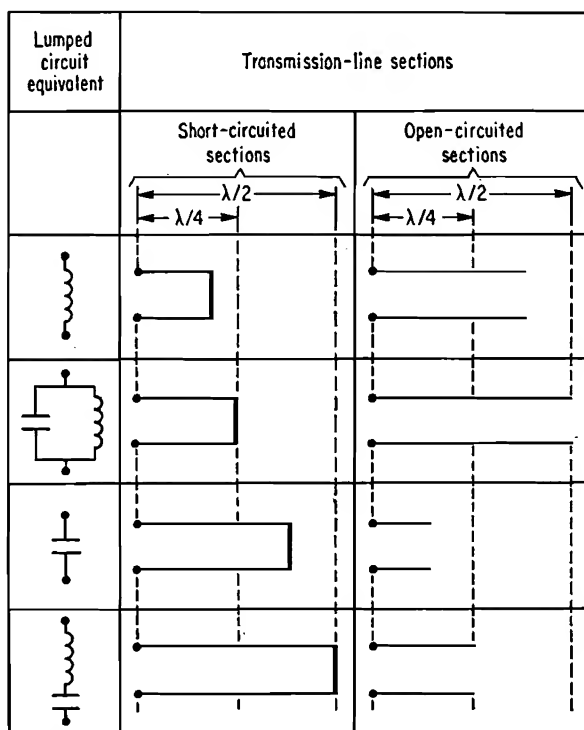


FIG. 5-22. LC lumped circuit equivalents of transmission line sections.

5-16. Impedance Matching in Transmission Lines. At the vhf and uhf frequencies, lumped circuit components of the type used at the lower frequencies becomes impractical. Therefore, at these higher frequencies, it is common to use reactances formed by sections of transmission line for the impedance-matching elements. Actually, the impedance-matching configurations described in previous sections can be formed in this manner. Figures 5-22 and 5-23 show the lumped circuit equivalents of some transmission-line sections which can be used to form the T , L , and π networks previously described.


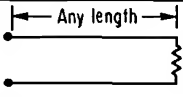

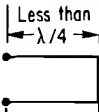
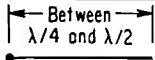

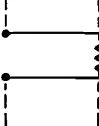
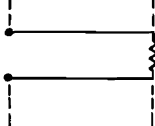


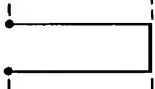
Lumped circuit equivalent	Transmission-line sections	
	 Any length Load resistor is equal to transmission-line impedance	
	 Less than $\lambda/4$	 Between $\lambda/4$ and $\lambda/2$
	 Load resistor is less than line impedance	 Load resistor is higher than line impedance
	 Load resistor is higher than line impedance	 Load resistor is less than line impedance

FIG. 5-23. LCR lumped circuit equivalents to transmission-line sections.

Sections of transmission line used for impedance-matching purposes are called *stubs*. These stubs may be of either the open-circuited or the short-circuited type, although the short-circuited type is generally preferred in most applications.

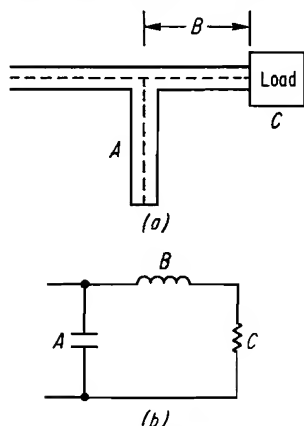


FIG. 5-24. (a) Single-stub tuner and (b) its equivalent lumped-circuit L network.

In practice, the identity of the transmission-line section is somewhat obscured because of the lack of any physical resemblance between it and its lumped circuit equivalent. Some of the more commonly used transmission-line coupling methods are described in the following sections.

The Single-stub Tuner. The single-stub tuner, when connected to a transmission line is the equivalent of an L network discussed in Sec. 5-8. This is illustrated in Fig. 5-24. The single-stub tuner is theoretically capable of matching any load im-

pedance to the characteristic impedance of the transmission line provided that the load is neither open- nor short-circuited and provided that the load does not consist of a pure reactance. To be able to match this wide range of impedances, the stub must be capable of being moved along the transmission line, as well as being adjustable in length.

The Double-stub Tuner. In many applications, moving a stub along a transmission line, as is required in the single-stub tuner, is inconvenient. The double-stub tuner, illustrated in Fig. 5-25, utilizes two adjustable stubs, fixed along a transmission line. The distance between the stubs

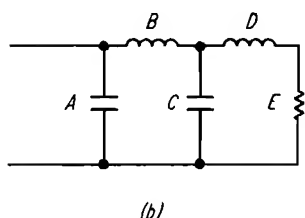
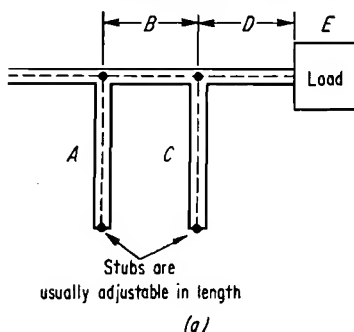


FIG. 5-25. (a) Double-stub tuner and (b) its equivalent lumped-circuit π network.

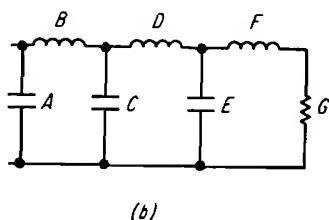
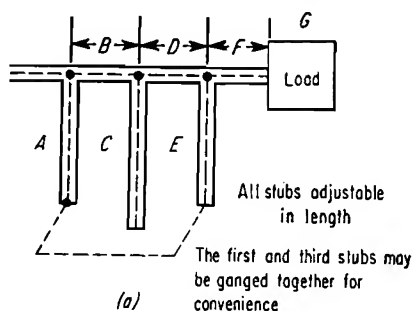


FIG. 5-26. (a) Triple-stub tuner and (b) its equivalent lumped-circuit network.

is a compromise between the ease of obtaining the match and the impedance-matching range. Increasing the distance between the stubs toward a half-wavelength increases the impedance range of the tuner, but it also increases the criticalness of the adjustment. As a compromise, the distance between the stubs is typically made a multiple of an odd quarter wavelength. With this spacing, the double-stub tuner is capable of matching a resistive load at the stub closest to the load no more than one-half the characteristic impedance of the line. Adjusting the length between the tuner and the load enables the tuner to match any impedance to the line.

The Triple-stub Tuner. A triple-stub tuner is theoretically capable of matching an impedance to any other impedance. The general construc-

tion of such a tuner is illustrated in Fig. 5-26. The three stubs are placed one-quarter wavelength apart on the transmission line. Unlike the double-stub tuner, the triple-stub tuner can be placed at any arbitrary distance from the load and still achieve a satisfactory match. For convenience in adjusting the triple-stub tuner, the first and third stubs can be ganged together. This reduces the number of adjustable variables down to two.

5-17. Impedance Matching in Waveguide Circuits. Much of the discussion of Sec. 5-16 concerning impedance matching in transmission lines is also applicable to impedance matching in waveguide circuits.

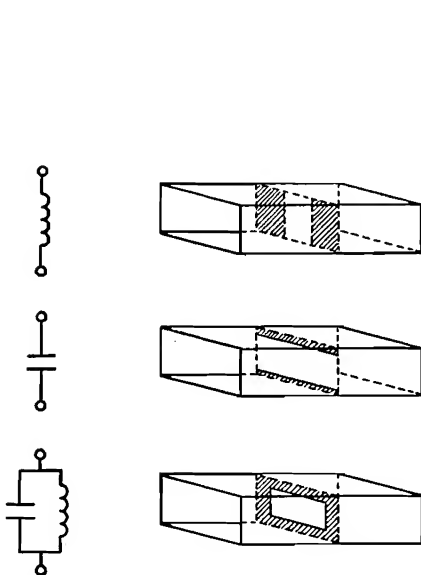


FIG. 5-27. Lumped-circuit waveguide equivalents.

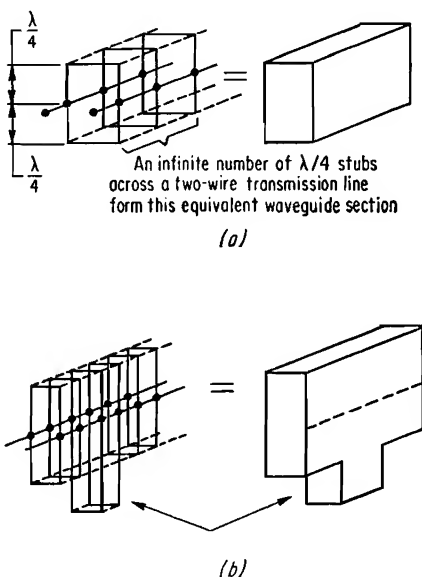


FIG. 5-28. The development of a waveguide stub from open-wire transmission line.

Waveguide sections also have their lumped circuit equivalents; however, these equivalent circuits depend on the field configuration or mode that is being propagated in the waveguide section. These waveguide modes, which are classified according to the configuration of the magnetic or electric field, are discussed further in Chap. 11. However, the lumped-circuit waveguide equivalents shown in Fig. 5-27 are known as the $TE_{1,0}$ mode for the most commonly used mode. This is the mode in which only one voltage maximum is present on each of the long sides of the waveguide, and none is present on the short side.

In addition, the transmission-line equivalents of Figs. 5-22 and 5-23 are also applicable to waveguide if the fundamental concept of wave-

guide development is considered. Since a waveguide is fundamentally an infinite number of quarter-wavelength stubs connected as shown in Fig. 5-28a, then any stub connected across a transmission line as shown in Fig. 5-28b will act according to the characteristics depicted for that stub in Figs. 5-22 and 5-23. The waveguide equivalent is also shown in Fig. 5-28b.

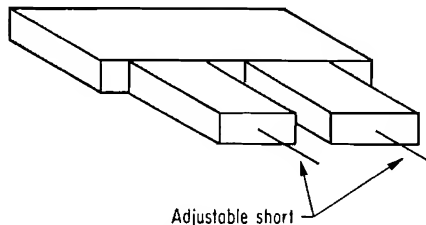


FIG. 5-29. A waveguide double-stub tuner.

Thus the waveguide equivalent of the coaxial transmission-line double-stub tuner previously shown in Fig. 5-25 can be constructed as shown in Fig. 5-29.

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AMPLITUDE MODULATION

6-1. Amplitude Modulation. The process of amplitude modulation (1,2) is quite similar to the process of mixing which was discussed in Sec. 2-18. Other terms which refer to the same process are *frequency changing*, *frequency translation*, *beating*, *heterodyning*, and *conversion*. In all cases, two frequencies are applied to a circuit, and at the output there appear the sum and difference of the two frequencies in addition to the original frequencies. Very seldom do all four frequencies appear at the output with any useful amplitude.

In a modulated r-f amplifier, the radio frequency and the sum and difference of the modulating and radio frequencies appear at the output. The impedance of the r-f output circuit is generally very low at the modulating frequencies, and the output therefore will not contain the modulating frequency.

6-2. Sidebands. If the voltage appearing at the output terminal of a radio transmitter is plotted against *time* or displayed on an oscilloscope with a linear time base, a continuous wave with varying amplitude or phase depending on the type of modulation will be observed. At any instant, the output voltage can have only one value, no matter how complicated the modulation may be. In the case of pulse modulation, the amplitude will periodically drop to zero. If, on the other hand, the voltage output is plotted or displayed against a linear *frequency* base, as can be done with an instrument known as a *spectrum analyzer* (3), the carrier frequency and the side frequencies or *sidebands* (4) will appear as vertical lines with heights dependent on the type and degree of modulation. This type of display generally yields much more information than the linear-time base observation.

Sidebands Produced by Amplitude Modulation. When a radio-frequency is 100 per cent modulated, the output signal will be as shown in Fig. 6-1*a* against a time base and as in Fig. 6-1*b* against a frequency base. It will be seen that at 100 per cent amplitude modulation, there will be two sidebands, each one-half the amplitude of the carrier and spaced away from the carrier by a frequency equal to the modulating frequency. A number of modifications to this type of modulation can be performed by filtering or other methods. These include

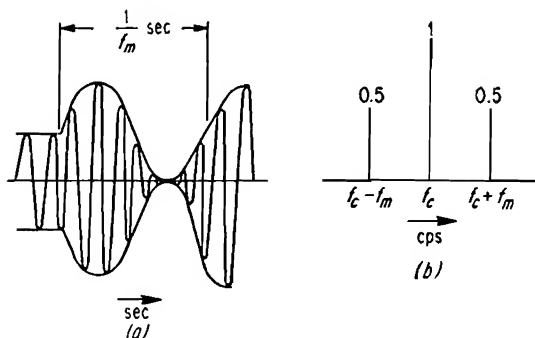


FIG. 6-1. Single-tone 100 per cent a-m DSB modulation. (a) Amplitude vs. time; (b) amplitude vs. frequency.

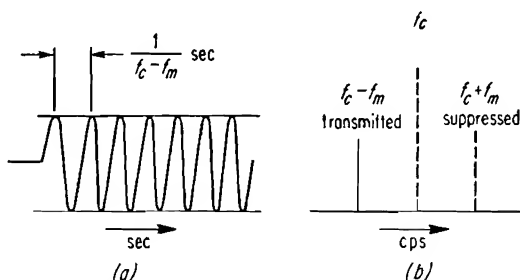


FIG. 6-2. Single-sideband modulation, single tone. (a) Amplitude vs. time; (b) amplitude vs. frequency.

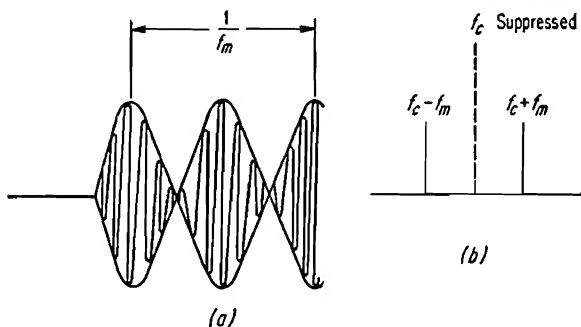


FIG. 6-3. Double-sideband suppressed carrier, single tone. (a) Amplitude vs. time; (b) amplitude vs. frequency.

1. *Single-sideband.* In this case, the carrier and one sideband are completely eliminated, as in Fig. 6-2a and b. Methods of ssb generation are discussed later in this chapter.

2. *Double-sideband, reduced carrier.* Here the carrier amplitude is reduced to save power and is inserted at the receiver (see Fig. 6-3a and b).

3. *Vestigial-sideband.* This type of transmission is generally used in television to conserve bandwidth. In American television the carrier

and upper sidebands are not reduced at all; but the lower sidebands are attenuated starting at 0.75 Mc below the carrier and are at least 20 db down at 1.25 Mc and more below the carrier. For frequencies up to 0.75 Mc the modulation displays will be as shown in Fig. 6-1, and for frequencies above 1.25 Mc the display will be as shown in Fig. 6-4a and b.

6-3. Double-sideband Amplitude Modulation. Amplitude modulation without sideband restriction was the first method used for modulating radio transmitters, and for many applications continues to be the most practical method. The choice of the "best" type of modulation is discussed in Chap. 1. The simplest means of amplitude modulation

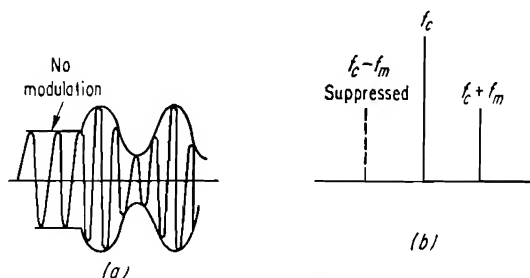


FIG. 6-4. Vestigial-sideband modulation, single tone. (a) Amplitude vs. time; (b) amplitude vs. frequency.

is merely to switch the carrier on and off. This is termed *on-off keying* and is often called c-w, although c-w strictly means unmodulated continuous wave.

The expression for a dsb full-carrier a-m wave is

$$e = E_c \sin 2\pi f_c t + \frac{mE_c}{2} \sin 2\pi(f_c - f_m)t + \frac{mE_c}{2} \sin 2\pi(f_c + f_m)t \quad (6-1)$$

where e = voltage at any time t

E_c = maximum amplitude of unmodulated signal

m = degree of modulation as defined by

$$\text{Positive peak } m = \frac{E_{\max} - E_c}{E_c} \quad (6-2)$$

$$\text{Negative peak } m = \frac{E_c - E_{\min}}{E_c} \quad (6-3)$$

E_{\max} = amplitude at positive peak of modulation

E_{\min} = amplitude at negative peak of modulation

f_c = carrier frequency

f_m = modulating frequency

The last two terms in Eq. (6-1) represent the sidebands.

There are four general ways of achieving amplitude modulation:

1. An r-f signal is passed through a circuit in which the r-f level is controlled by the modulating voltage. The diode modulator is an example.
2. An r-f signal drives an amplifier which has constant efficiency over a range of the d-c input voltage. The modulating voltage varies the d-c input voltage. The plate-modulated amplifier is an example.
3. An r-f signal drives an amplifier which has constant d-c input voltage but has its efficiency and input (d-c) current varied by a modulating voltage. A grid-modulated amplifier is an example.
4. Two phase-modulated transmitters, driven by a common r-f generator and connected to a common load, are simultaneously phase-modulated so that the combined output varies from zero to twice the unmodulated amplitude. This is called *phase-to-amplitude* or *outphasing* modulation.

6-4. On-off Keying. On-off keying has largely been replaced by *frequency shift keying* (fsk) for long-distance high-speed communication circuits (see Chap. 12). This is because the signal-to-noise ratio at the receiver may be about 10 db worse for frequency-shift keying than for equivalent performance from an on-off keyed signal. On-off keying, remains, however, the principal means of communication between maritime mobile stations and coast stations.

On-off keying is usually accomplished at low power levels so that the power that need be controlled is small. The amplifiers following the keyed stage must be designed so that the plate currents will drop to zero or to a safe value when r-f drive is cut off. In certain types of operation the operator may be receiving and transmitting on the same frequency and may wish to "listen through" so that the operator at the other end can "break in" while the former is sending. This is one reason why frequency-shift keying is not used on manually operated shipboard transmitters.

Where "break-in" is needed it may be necessary to key the transmitter oscillator to avoid generating an interfering signal or *backwave* in the receiver. In general, however, it is undesirable to key the oscillator because it is much more difficult to eliminate undesirable keying characteristics such as *key clicks*, *chirp*, or *lilt* on the signal. One method of avoiding key-up oscillator interference is to use a mixing system in which the output frequency is the sum or difference of two oscillators (5). The mixer can be keyed to stop generation of the output frequency. The individual oscillators would still have to be well shielded to avoid generation of an interfering signal by mixing in the first stage of the receiver. Another method is to operate the oscillator at a submultiple of the output frequency and design the oscillator so that it has negligible harmonic output (6).

Methods of Keying. The simplest method of keying is to break the cathode circuit in the stage being keyed. When economy prevents the use of a negative bias supply to cut off amplifiers following the keyed stage when the drive is removed, a *clamp tube* may be used to reduce the screen voltage. This circuit was discussed in Sec. 3-19. When a negative bias supply is available, amplifier tubes or the oscillator itself, may be cut off directly with high grid bias. This is called *blocked-grid keying*. Alternatively an additional *keyer tube* may be used. This can be used as an electronic switch which will be connected in series with the circuit to be keyed. High bias will be applied to the grid of the keyer tube with the key open, and this will be reduced when the key is closed.

Keying Transients and Chirps. The following problems must be considered in the design of a radiotelegraph transmitter:

1. Frequency variation or chirp
2. Local key clicks due to transient voltages generated in the keying control circuits
3. Transmitted key clicks
4. Power-supply variations

Frequency variation of a keyed oscillator (7) can be reduced by making the oscillator frequency independent of supply voltages. This can be done by regulating the voltage applied to the oscillator or by using oscillator circuits which are isolated from the associated tube (see Sec. 2-2).

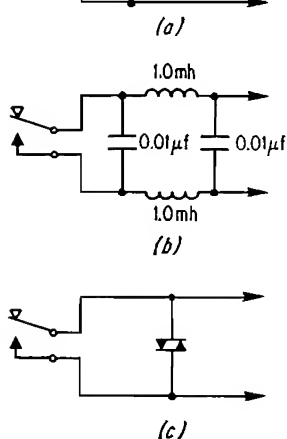


FIG. 6-5. Key-click filters.

Elimination of local key clicks is usually a simple matter. Figure 6-5a, b shows two click filter circuits. If the key is operating an inductive circuit such as a relay coil, large voltages may be generated when the circuit is broken. These can be reduced with special diodes using the circuit shown in Fig. 6-5c (8).

Elimination of transmitted key clicks may not be possible with simple filters at the key contacts. The principal cause of clicks is too rapid a rise or fall in the *envelope* of the transmitted signal. For example, if the envelope rises from zero to full value in 10 μsec, the leading edge of the signal would have the same shape as one modulation cycle of an r-f signal modulated with a frequency of 100 kc. Sidebands 100 kc on each side of the carrier would therefore be expected. There is no need for the rise time to be any faster than that sufficient to transmit the third harmonic

of the keying frequency. At 100 words per min the rise time can be as slow as 5 msec. Inductance in series with the key (9) will delay rise time, and shunt capacity across the key will delay fall time, but such treatment may be useless if the keyed stage is overdriving the following stage. Also, if a crystal oscillator is being keyed, the crystal may not oscillate at all until the voltage has reached an appreciable value, and then the oscillator output voltage may rise very rapidly to a value sufficient to drive the following stage to full output (10,11). In these cases the key-click filter would delay the start of the signal but would not do much to "round off" the leading edge. An exciter, which is itself quite free of clicks, may cause severe clicks when used to drive an overbiased amplifier (12). The rise and fall times of the output envelope, in addition to spurious sidebands generated in lower-level stages, must be examined in studying a key-click problem.

Another cause of severe clicks is instability in an amplifier. Parasitic oscillations may be triggered by the keying but may stop when the amplifier is up to full drive. Radiation of key clicks may be confined to a narrow band around the transmitted frequency if sufficiently selective r-f output circuits are used (13). For example, a π network has good high-frequency attenuation but has little effect at frequencies below the carrier, and it is possible for communication transmitters to radiate key clicks throughout the 550- to 1,600-kc broadcast band when a transmitter generates excessive clicks and has a π network in the output.

Power-supply Transients. The fractional dip or rise in the output voltage of a power supply when a resistance R is placed at the output of a single stage LC filter is

$$\frac{\Delta E_0}{E_0} = \frac{\sqrt{L/C}}{R} \quad (6-4)$$

and the time required to reach the peak or dip is

$$T = \frac{\pi}{2} \sqrt{LC} \quad (6-5)$$

Equation (6-4) shows that the rises and dips will be reduced as L is reduced and as C is increased. This problem is avoided when frequency-shift keying is used because the load on the power supply is then constant. The problem may also be avoided by the use of an *absorber*, which is a circuit consisting of a keyer tube and a resistive load. When the key is up, the keyer tube conducts transferring the power-supply output to the load, thus keeping the load on the power supply constant.

6-5. Diode Modulator. When an audio frequency and a radio frequency are combined in a *linear* device, such as a transformer, modulation does *not* take place and the two waves will merely be *superimposed*,

and the result, shown on a time base, will be as in Fig. 6-6. Notice the difference between this waveform and the one shown in Fig. 6-1a. A linear circuit or device is one in which the output is at all times proportional to the input. A transformer becomes *nonlinear* when an excessive voltage is applied to the primary because the iron will *saturate*, and the secondary voltage will stop rising as the primary voltage is further increased. Similarly a resistor is linear as long as it obeys Ohm's law; but, if a carbon resistor is overloaded, the heat will increase the value of the resistance, and an increase in voltage will no longer result in an exactly proportional increase in current. An example of a linear device is a high-fidelity sound system. If two audio

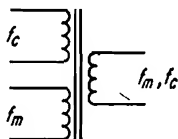
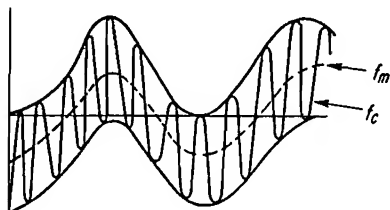


FIG. 6-6. Linear adding circuit.

tones are fed into the amplifier, the motion of the speaker diaphragm will be as represented in Fig. 6-6, and no other frequencies will be generated.

An example of a nonlinear device is a diode rectifier. One polarity of signals will be transmitted without change, but the opposite polarity will be stopped. A diode modulator (14) is shown in Fig. 6-7. If the wave shown in Fig. 6-6 is applied to the diode, the output will have the

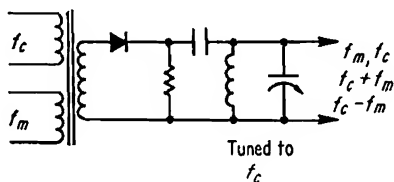


FIG. 6-7. Diode modulator.

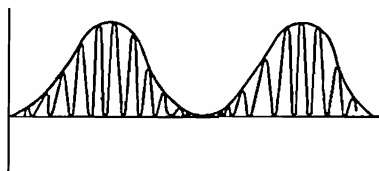


FIG. 6-8. Waveform clipped by action of diode.

negative half clipped off, and it will be as shown in Fig. 6-8. If the output of the diode is connected to a tuned circuit tuned to f_c , the *flywheel* effect of the LC circuit will insert the negative half of each r-f cycle, and the envelope will be a mirror image of the top and will look like Fig. 6-1b rather than Fig. 6-6. Vacuum tubes and transistors may also be used as nonlinear elements. All mixers, converters, frequency changers, and diode modulators are nonlinear circuits and behave in this manner. If they are not nonlinear, the sum and difference products will not be generated. Most sound-reproducing systems are more or less nonlinear,

and the sum and difference frequencies which result are called *inter-modulation* products or distortion.

The diode modulator shown in Fig. 6-7 is only useful for very low levels, and it would introduce considerable *insertion loss*. Consequently, it will not be found in a radio transmitter for ordinary a-m purposes. Diode-balanced modulators of the bridge or ring type are used, however, for the purpose of generating double-sideband suppressed-carrier signals. For such use the carrier voltage is usually 10 to 20 times the peak modulating voltage (15).

6-6. Plate Modulation. When the modulating voltage is superimposed on the d-c plate supply for an amplifier, it is said to be *plate-modulated*. This is the most widely used method of obtaining amplitude modulation. Plate modulation is generally applied to an amplifier which is adjusted to operate in the class C condition (see Sec. 3-3). When a triode tube is plate-modulated, the following points must be considered in selecting operating conditions. Additional limitations are imposed by the screen grid in a tetrode, and these will be discussed later.

1. The r-f drive voltage, output load impedance, and grid bias should be adjusted so that at the peak of the modulation cycle, when the plate supply voltage is $2E_b$, the peak r-f output voltage will be twice what it was when the plate voltage was E_b . The peak plate-to-cathode voltage should be almost as much as $2E_b$ at the peak of the modulation cycle but can never quite reach this value because the minimum plate voltage E_{\min} should not fall below the peak positive-grid voltage.

2. If the r-f drive is enough to produce *saturation* at the peak of the modulation cycle, it will be more than enough through the rest of the cycle. Saturation is the condition reached when further increase in r-f drive produces no more r-f output. At the modulation trough, the r-f drive may cause large grid dissipation unless precautions are taken.

The objective in adjusting a plate-modulated class C amplifier is to obtain as nearly a linear relationship as possible between the modulating voltage and the envelope of the modulated r-f wave (16). Figure 6-9 shows the r-f output voltage relationship to the plate-supply voltage in a typical class C amplifier (17). The linear region is that in which the r-f drive is enough to saturate the amplifier. For optimum operation, the unmodulated plate voltage E_b is selected halfway along the linear region. For 100 per cent modulation, the modulating voltage e_m will be equal to E_b .

If the modulating voltage swings the peak r-f plate-cathode voltage into the nonlinear or unsaturated region, the r-f voltage will not rise so much on the positive peaks as it swings down on the negative peaks. In the extreme condition where there is only a small positive rise and a large negative drop, the condition is known as *downward modulation*. This

condition may also be called *carrier shift*. Downward modulation is an allowable type of operation for type A2 modulated telegraphy but will cause severe distortion in type A3 telephony operation.

In order to avoid damage to the grid by excessive excitation at the trough of the modulating cycle, the driver should have poor *regulation*, which means that it has barely sufficient power to drive the grid when the grid current is at a minimum at the peak of the modulation cycle. This is opposite to the requirement for driving a class B linear amplifier which

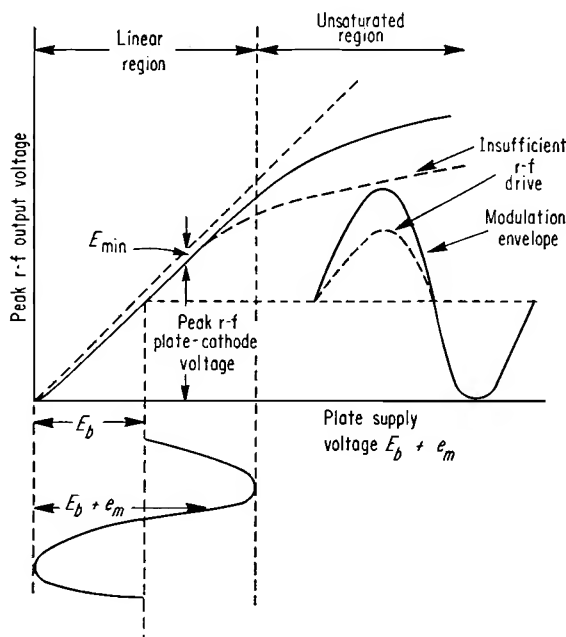


FIG. 6-9. Relationship between radio-frequency output voltage and modulating voltage in a typical plate-modulated class C amplifier.

requires a driver with good regulation. The grid of a class C amplifier can be automatically protected by obtaining at least part of the grid bias from a series grid resistor. When the grid current tends to rise, the bias will rise and reduce the grid current. The capacitor shunting the grid resistance should have a reactance that is at least twice the grid resistance at the highest modulating frequency; otherwise the bias will not follow the modulating cycle. It is desirable to have sufficient fixed grid bias in series with the grid resistor to prevent damage to the amplifier tube in the event the r-f drive fails. The use of grid-resistance bias improves the linearity of the dynamic characteristic shown in Fig. 6-9 and also increases the power output over what would be obtained with

fixed or cathode bias. An exact graphical analysis of the performance of a self-biased modulated class C amplifier is a complicated procedure because the grid bias is constantly changing. A procedure for doing this has been given by Sarbacher (18).

Power Requirements of the Modulator. As can be seen from Fig. 6-1b, the power in the sidebands of a 100 per cent modulated amplifier is 50 per cent of the carrier power. This additional power must be supplied by the modulator. The available modulator power must therefore be 50 per cent of the unmodulated *input* to the amplifier. For modulation depths of less than 100 per cent the modulator power will be proportional to m^2 . The modulator must deliver its power to a load impedance which consists of the combination of

1. Resistive load which is equal to E_b/I_b
2. Capacitive load which is equal to the plate bypass capacitor plus any other capacity to ground appearing at the d-c input terminal on the modulated amplifier

Modulation of Tetrode Amplifiers. When amplifier tubes which have a screen-grid (tetrodes, beam tetrodes, and pentodes) are modulated, it is necessary to develop about 70 per cent amplitude modulation of the screen voltage. If the screen voltage alone is modulated, this is a form of efficiency or grid modulation and will be discussed later. Modulation of the screen will require additional power, but this will be supplied by the screen supply and not by the modulator. The ratio of the screen-modulating power to plate-modulating power is the same as the ratio of screen-to-plate current and varies between 0.05 and 0.15. It is important that the instantaneous plate voltage does not fall below the instantaneous screen voltage because excessive screen current will flow under this condition. The screen voltage must therefore be reduced in synchronism with the plate voltage.

Modulation of the screen may be developed in one of the following ways (19):

1. The screen voltage can be supplied through a dropping resistor connected to the unmodulated d-c plate supply.
2. When a low voltage, fixed-screen-voltage supply is used, a modulation choke is placed in series with the supply. In the case of voice modulation a value of 10 henrys is satisfactory.
3. A third winding can be placed on the modulation transformer designed to develop the required screen modulation voltage.

These three methods are shown in Fig. 6-10. It is important that the bypass capacitor at the screen is a low impedance to the radio frequency but a high impedance to the audio frequency; otherwise the screen will not follow the modulation.

Heising or Constant-current Modulation. Figure 6-11 shows one of the oldest types of modulators (20). This is the Heising method which has poor efficiency because the modulator must operate as class A. The load impedance presented by the class C amplifier should load the modulator correctly. For 100 per cent

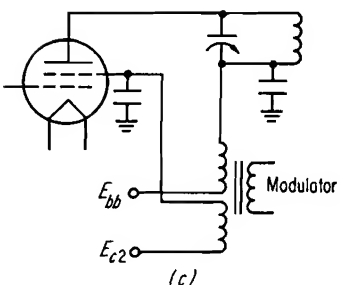
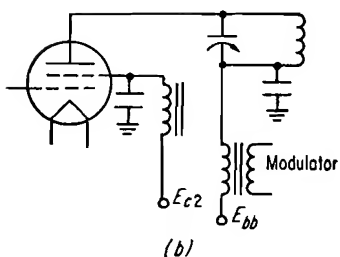
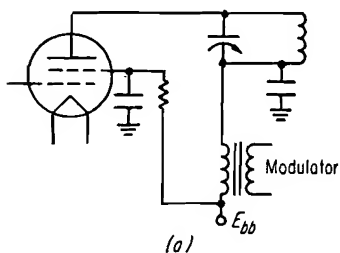


FIG. 6-10. Methods of simultaneously modulating the plate and screen. (a) Series resistance; (b) series inductance; (c) screen modulation winding on transformer. (Courtesy of Eitel-McCullough, Inc.)

modulation, the plate voltage on the modulator must be higher than the plate voltage on the r-f amplifier. This is because the voltage developed by the modulator cannot swing to zero without a great deal of distortion. This is accomplished by inserting the resistance R between the modulator and r-f amplifier. This must be bypassed by C , which should have a reactance at the lowest modulating frequency which is not more than one-tenth the value of R . The voltage drop across R must be equal to or more than the minimum instantaneous voltage which will appear at the modulator output. Under ideal operating conditions the rise or fall in the current taken by the r-f amplifier will be just equal to the fall or rise in the current taken by the modulator, and the current taken by the modulator will be a constant. The modulator tube must be adjusted to operate as a class A amplifier which normally has poor efficiency, and when the loss in the resistor R is considered, it will be seen that this is a low-efficiency system.

Class AB and Class B Modulators. Efficient modulators can be built by using two tubes in a push-pull circuit and coupling to the r-f amplifier with an impedance-matching transformer. Efficient operation of the tubes is achieved by biasing them so that they are cut off for up to

half of the time. One tube amplifies the positive half cycles, and the other tube amplifies the negative half cycles. The transformer combines the half cycles to give an approximately undistorted output. The maximum theoretical efficiency of a class B modulator is 78.5 per cent, but the practical efficiencies are closer to 60 per cent. See Chap. 3 for a discus-

sion of class AB and B amplifiers. This discussion deals with r-f tuned amplifiers; but the performance with audio is the same except that two tubes in push-pull must be used in the latter case. In the tuned r-f amplifier the output wave is restored to its sinusoidal form by the "fly-wheel" effect of the tuned circuit.

In class B operation, the power input to the modulator is proportional to the signal, and, when there is no audio input, the plate current will be low or zero. This fact results in an important improvement in efficiency when the input is speech which has a small average amplitude compared with the peaks. The distortion in a class B modulator may be improved

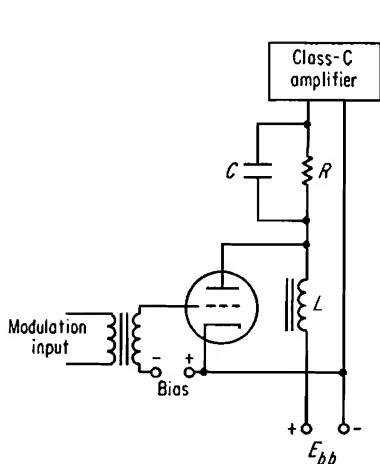


FIG. 6-11. Heising or constant-current modulator.

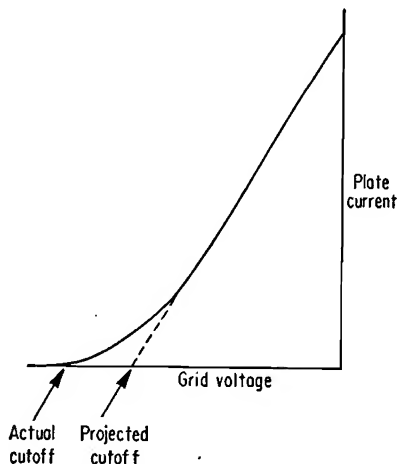


FIG. 6-12. Projected cutoff bias.

by adjusting the bias for what is called *projected cutoff bias*. The grid-voltage-plate-current curves of typical tubes are not linear near the cutoff bias, and less distortion is obtained if the bias is adjusted for the point where the cutoff would be if the characteristic were straight. This is illustrated in Fig. 6-12. If the two tubes have the same characteristic, the curved portions will cancel, and an over-all straight characteristic for the two tubes can be obtained.

It is important that the bias supply for a class B or class AB modulator be fixed. Cathode bias cannot be used because the cathode current is not constant. In the case of a tetrode class B modulator, the screen voltage must also be fixed. A screen-dropping resistor must not be used.

Modulator Driver Stage. In order to obtain optimum performance from a class B modulator, it is usually necessary to drive the grids to grid current. This means that power must be supplied. In the case of the class AB₁ modulator, no grid driving power is required, and in

many cases the simplification of the driver offsets the poorer efficiency of the class AB_1 output stage. The grids of a class B modulator represent a variable load resistance over the a-f cycle because the grid current is not directly proportional to the audio voltage. This means that the driver stage must have good regulation. This can be achieved by using triodes with a low plate resistance or by using *feedback* to reduce the apparent plate resistance of tetrodes or pentodes. It is usual practice for the driver stage to be coupled to the class B modulator grids through a step-down transformer to improve the regulation.

Series or Constant-voltage Modulation. Another type of modulation (20) which is somewhat similar to the Heising circuit is the constant-voltage circuit. This has the advantage that no modulation transformer or choke is needed, and this may be important where weight must be kept down. A simplified circuit is shown in Fig. 6-13. This modulator must operate as a class A amplifier. The power-supply voltage is constant and the voltage division between the modulator and the r-f amplifier

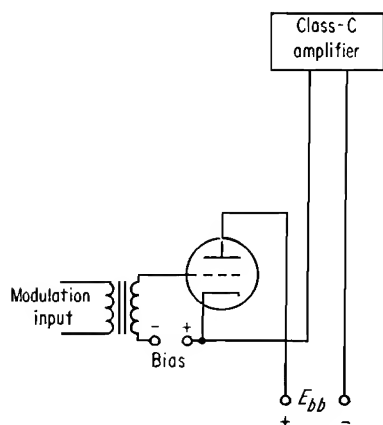


FIG. 6-13. Series or constant-voltage modulator. (Courtesy of George Neumes, Ltd.)

varies throughout the modulation cycle. The r-f amplifier is the plate load for the modulator. As with the constant-current modulator, the modulator plate voltage must be greater than the r-f amplifier plate voltage if 100 per cent modulation is to be achieved. The grid-bias voltages can be adjusted so that any desired voltage division is obtained for the unmodulated condition. One hundred per cent modulation in the negative direction cannot be achieved because this would reduce the plate current to zero in the r-f amplifier; since the modulator is in series, its plate current would also drop to zero

with consequent distortion. Another difficulty with this circuit is that either the r-f amplifier or the modulator must have its filament circuit insulated from ground by the supply voltage. The r-f amplifier is usually operated closer to ground.

6-7. Grid Modulation. The modulating voltage can be applied to one of the grids in an amplifier with a considerably smaller amount of power than is required for plate modulation. The available r-f carrier power will, however, be reduced. The operating principles are the same regardless of the grid that is used. The plate voltage remains constant and the increase in power output with modulation is obtained by making

both the plate current and plate efficiency vary with the modulating signal. In order to obtain four times the unmodulated carrier power with constant plate voltage, both the plate current and plate efficiency must be doubled. The usual output power is one-quarter to one-third that obtainable for plate modulation with an average tube. The unmodulated carrier efficiency is usually about 33 per cent.

Control-grid Modulation. The control grid of an r-f amplifier is modulated by inserting the modulating voltage in series with the grid-bias supply. The instantaneous grid bias will change with the modulating voltage, and this will change the period of time over which the r-f drive causes rectified grid current to flow. This places a variable load on both the r-f driver and the modulator. Because of this, both the r-f driver and the modulator should be constantly loaded with resistance in order to reduce distortion. Control-grid modulation is the most practical method of introducing video modulation to a television transmitter.

Modulating frequencies for video transmission are as high as 4.3 Mc. The grid-to-cathode capacitances and peak-to-peak video voltages required for 100 per cent modulation are both low enough in suitable transmitting tubes to permit practical video modulators to be built. With grounded-cathode operation of the grid-modulated stage, no sidebands will occur in the input circuit, provided there is no r-f feedback, and only the output circuit will need to be broadbanded. Tetrodes with high transconductances are suitable for grid-modulated video amplifiers. They should be neutralized if necessary to reduce output-input coupling to a minimum.

Typical operation of a grid-modulated tetrode is illustrated in Fig. 6-14 (21). In order to obtain a flat response over a 4.5-Mc range, a double-tuned circuit will probably be selected for the plate circuit, and the maximum load impedance that will permit this bandwidth will determine the slope of the load line AB . (Refer to Sec. 3-17 for a discussion of double-tuned circuits.) The grid bias corresponding to the maximum power output of the amplifier (synchronizing peak level in a video signal) is chosen to equal the cutoff bias of the tube, which is the point B . The r-f drive is then chosen to swing the plate current up to point A . The point A is chosen so that the swing does not extend into the crowded region of the tube characteristics, which means that the minimum plate voltage is not allowed to drop below the screen voltage.

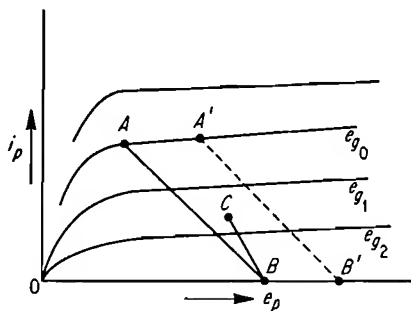


FIG. 6-14. Operation of a grid-bias modulated tetrode. (Courtesy of RCA.)

If the plate voltage drops below the screen voltage, the screen current and screen dissipation will rise sharply. The required peak-to-peak voltage swing at the modulator output will be approximately equal to the difference between the grid bias voltages corresponding to A and B .

The r-f drive should remain appreciably constant, and, at the negative peak of modulation, the r-f drive should just drive the tube into conduction. If the plate voltage is increased to B' , the load-line slope will not change, and the swing will be to point A' . The screen current will reduce while power output and plate dissipation will increase. Moving toward B will increase the efficiency, and moving toward B' will improve the linearity. The designer must choose a compromise.

Line CB illustrates the conditions during the time when the picture is gray. The slope changes because the angle of plate current flow becomes shorter.

Screen-grid Modulated Amplifiers. Screen-grid modulation is an alternative to control-grid modulation with the disadvantage that a bit more power is required. In addition, plate characteristics for varying values of screen voltage are not generally given in tube data sheets, and the performance must be determined experimentally. The amplifier is first adjusted so that at the peak of the modulation it is operating at optimum class C efficiency. The screen voltage for optimum efficiency is determined and then cut to half this value for the unmodulated condition. In some tubes, the plate current may not be cut off at zero screen voltage, and complete cutoff will require the screen to be swung negative. One advantage of modulating the screen is that the load presented by the screen is substantially independent of the applied voltage.

The screen grid may be modulated by the clamp-tube circuit shown in Fig. 3-39 (22). The clamp tube must be operated as a class A amplifier, and the operation is similar to the Heising or constant-current modulation system. As with the constant-current system the voltage applied to the clamp-tube plate must exceed that applied to the modulated amplifier screen if 100 per cent modulation is to be approached. This is usually done by connecting a bypassed resistor between the clamp-tube plate and the screen of the r-f amplifier. A method which is more efficient is to use a transformer to couple the modulated voltage to the screen.

Suppressor-grid Modulation. This is a third type of efficiency modulation which is not too common because pentodes with the suppressor grid brought out to a separate pin are comparatively rare. Suppressor-grid characteristics are not usually given in tube data, and there is the further difficulty that the manufacturers often do not control the suppressor-grid characteristics very closely. This means that tube substitution may require changes in the operating conditions. This is not a tolerable

commercial practice. Since the suppressor is always negative, the modulator is not required to furnish any power at all, and there is no problem of regulation. The method of adjustment is similar to that used for setting up a screen-grid modulator. The amplifier is adjusted for optimum class C operation with the suppressor grid at zero voltage, and then the suppressor grid bias is set to reduce the plate current to half. The peak modulating voltage will then be equal to this bias.

Cathode Modulation. Cathode modulation is a combination of plate and grid modulation and permits a carrier efficiency which is midway between the two. The unmodulated carrier efficiency will be lower than the efficiency at the modulation peak by an amount which depends on the proportion of grid-to-plate modulation.

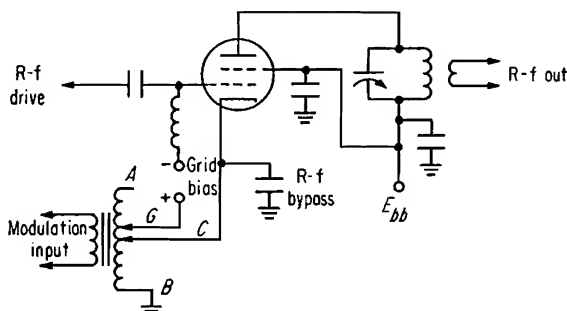


FIG. 6-15. Cathode modulation. (Courtesy of ARRL.)

The basic cathode modulation circuit is shown in Fig. 6-15. The relative plate and grid modulations are determined by the relative positions of the taps on the modulation transformer. If tap *G* is all the way down to *B* and tap *C* is all the way up to *A*, the effect will be the same as if the transformer were in series with the positive supply lead, and this will produce 100 per cent plate modulation and no grid modulation. If the position of the taps is reversed, with *G* at *A* and *C* at *B*, the result will be no plate modulation and all grid modulation. The voltage and power required from the modulator will vary accordingly. The load impedance presented to the modulator will be given by (23)

$$R_L = m \frac{E_b}{I_b} \quad (6-6)$$

where m = degree of plate modulation

E_b = unmodulated plate voltage

I_b = unmodulated plate current

(E_b and I_b will remain constant during modulation if there is no distortion.)

Approximate values of d-c power input, r-f power output, required audio power, and plate efficiency for variation in m are given in Fig. 6-16. When high values of m are used, some grid-leak bias should be used in

addition to fixed bias for the reason given in the discussion under plate modulation.

Amplitude Modulation of High-powered Klystrons. When an a-m output is required from a klystron amplifier, it is usually operated as a linear amplifier, and modulation is applied to the driver stage. Klystrons

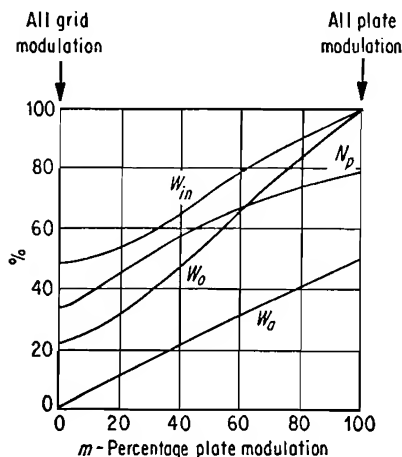


FIG. 6-16. Cathode-modulation performance curves, in terms of percentage of plate modulation plotted against percentage of class-C telephony tube ratings. (Courtesy of ARRL.)

W_{in} = d-c plate input watts in terms of percentage of plate-modulation rating

W_o = Carrier output watts in percentage of plate-modulation rating (based on plate efficiency of 77.5 per cent)

W_a = Audio power in percentage of d-c watts input

N_p = Plate efficiency of the amplifier in percentage

have high power gain, and an output of 10 kw with less than 10 watts of drive power is typical. When the modulation duty cycle is low as for single-channel voice and where it is desirable to keep power consumption at a minimum, the beam current of a klystron may be modulated with a *modulating anode* (24). Klystrons which are fitted with such anodes are designated by the inclusion of the letter *M* in the type number. The modulating anode has no voltage gain, and the peak-to-peak swing must be approximately equal to the beam voltage. The power required by the anode is negligible, but the modulator must have the power capability to develop the necessary voltage across the capacity of the modulating anode. This capacity is approximately 40 μf . This type of modulation is somewhat similar to grid or efficiency modulation. At the peak of the modulation cycle, the beam current and the efficiency are both approximately doubled from the unmodulated condition.

In the case of a vacuum-tube amplifier, the efficiency for class B linear operation is about the same as for grid modulation. In a klystron amplifier, the beam current is not at all related to the r-f drive. In order to obtain the peak r-f power output required at the peak of the modulation cycle, the beam current must be sufficiently high, and it must remain at this value throughout the modulation cycle. When the modulating anode is used, the beam current will instantaneously ride up and down with the modulation. Therefore, it can be seen that, unlike vacuum-tube amplifiers, the efficiency for linear operation is much less than for modulated-anode operation.

The modulator for the modulating anode is usually operated as a single-ended class A amplifier. Figure 6-17 shows a modulator for the Eimac type 3KM50,000PA klystron. This uses two triodes in a *cascode* circuit. The inductances resonate the modulating-anode capacity at the center of the desired passband. The RLC combination in the cathode circuit of the modulator provides controlled feedback, which improves the frequency response of the circuit.

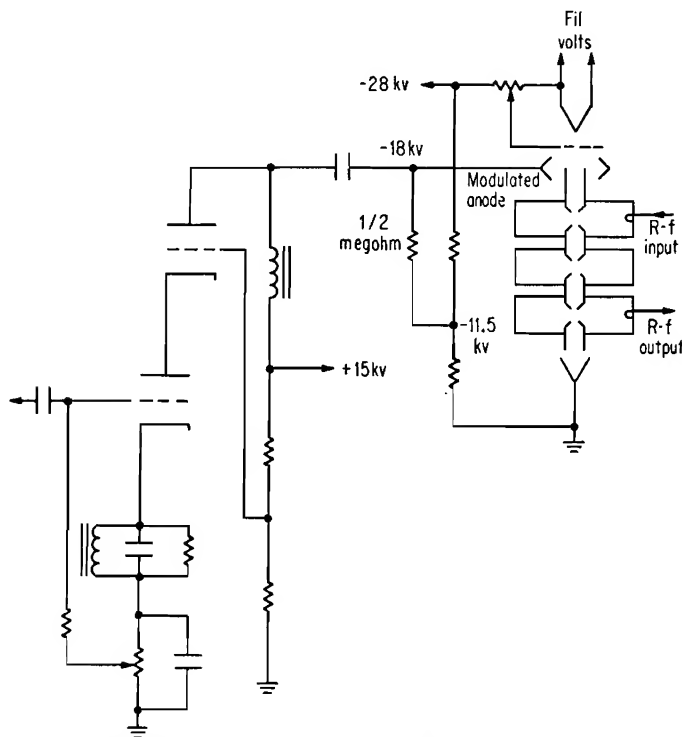


FIG. 6-17. Modulation of klystron modulating anode. (Courtesy of ITT.)

Controlled-carrier Modulation. In a low duty-cycle system, such as speech transmission, the r-f amplifier losses can be considerably reduced by the use of *controlled-carrier* circuits. A voltage is generated which is proportional to the average modulation level, and this in turn controls the amplitude of the carrier. The action must be slower than the lowest audio frequency. At the receiver, the automatic volume control must follow the level variations, and the detector must follow only the audio variations. The controlled-carrier circuits are essentially double modulators, one slow-operating, following the average modulation level, and the other following the audio (25-29). One convenient method is to

use a clamp tube as shown in Fig. 6-18 for the average level control. The rectifier in the modulator grid circuit provides the modulator with a d-c bias that varies with the average of the audio input level. As the audio

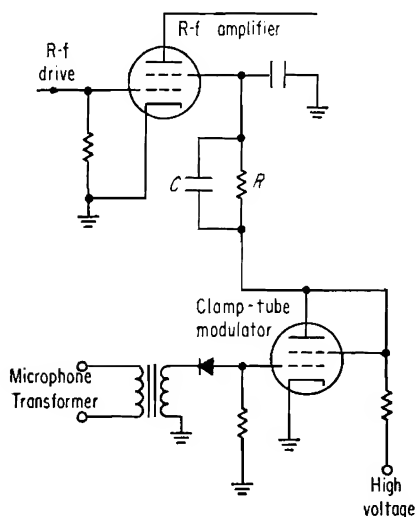


FIG. 6-18. Controlled-carrier modulation using a clamp tube. (Courtesy of ARRL.)

level increases, the bias on the modulator likewise increases. This reduces the modulator plate current, and thus the voltage drop through the modulator plate resistor. This allows the average or d-c voltage of the screen of the r-f amplifier to rise, and so the carrier level rises. The RC combination between the modulator and the r-f amplifier screen permits a higher level of modulation to be achieved. The circuit of Fig. 6-18 will also work if a capacitor is substituted for the rectifier such that the time constant of the RC circuit in the modulator grid is about 0.25 sec.

Single sideband is a much more satisfactory type of controlled r-f output signal because there is no r-f output at all when there is no modulation.

6-8. Amplification of Modulated R-F Voltages. A modulated signal can be amplified without distortion of the envelope, provided that the amplifier is operated in a linear manner. Class A, AB, and B but not class C amplifiers can be adjusted for linear operation. This is discussed in detail in Chap. 3. Klystron amplifiers are also fairly linear and can also amplify modulated signals. High-power transmitters generally employ linear amplifiers. The efficiency of a class B linear amplifier is only about 33 per cent at the carrier level, and, as efficiency is an important factor in a high-power amplifier, various efficiency-improving schemes have been used. Two such methods have been described by Doherty (30) and Fisher (31).

6-9. Carrier-suppression Techniques. Single-sideband suppressed-carrier techniques will be discussed later in this chapter.

It is also advantageous to reduce the carrier in dsb systems to improve efficiency. A certain level of reduction can be tolerated without the necessity for any special techniques to be used at the receiver. As the level of the carrier is further reduced, it becomes necessary to *exalt* or reinsert the carrier at the receiver. A simplified circuit for generating dsb reduced carrier is shown in Fig. 6-19. The r-f signal is applied to the

grids in push-pull, and the plates are connected in parallel. One screen is biased negatively, and the other is biased positively. The audio is applied to the screen grids in push-pull. This circuit can be adjusted so that the sideband power is much greater than the carrier power (32,33).

A somewhat similar system is called *supermodulation*. In this method (34-36) a signal is generated in which the carrier is reduced only during modulation. In this method, two grid-modulated amplifiers are connected to a common output circuit. A simplified schematic is shown in Fig. 6-20. Under the unmodulated carrier condition, one tube is biased beyond cutoff, and the other is delivering almost

its maximum output. When a modulating signal is applied, the output from the *carrier* tube drops, and the output from the *peak* tube increases. The operation of the peak tube causes the voltage at the plate of the carrier tube to be reduced. The carrier efficiency of this

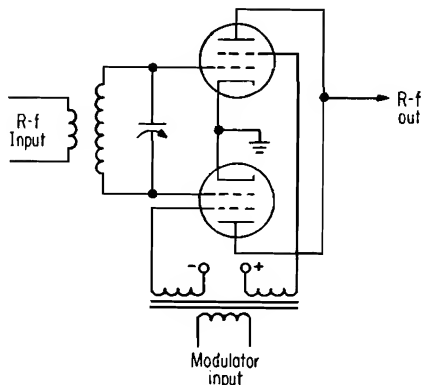


FIG. 6-19. Simplified dsbrc circuit. (Courtesy of ARRL.)

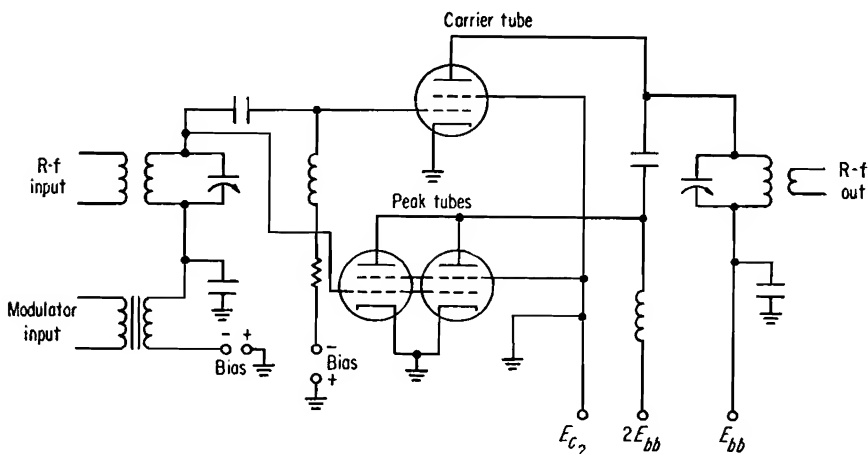


FIG. 6-20. Supermodulation circuit. (Courtesy of ARRL.)

circuit is about 66 per cent which is about twice that of a conventional grid-modulated stage.

6-10. Vestigial-sideband Transmission. In vestigial-sideband transmission, the carrier and upper sidebands are not attenuated while the

lower sidebands below a certain frequency are cut off. In the standard television transmission used in the United States the lower sidebands down to about 750 kc below the carrier are transmitted, and then they are rapidly attenuated so that they are at least 20 db down at 1.25 Mc below the carrier. For color-television transmitters the signal at 3.58 Mc below the carrier must be attenuated by at least 42 db. This type of transmission is used only to conserve bandwidth, and it does little to conserve power. In the 525-line system of transmitting television, the vestigial-sideband bandwidth is 6 Mc including the sound channel, as compared with the 9-Mc bandwidth that would be necessary for dsb transmission.

This type of transmission is generated simply by the use of passive filters in the r-f amplifiers or in the transmission line following the final output stage. Some advantage is taken of the fact that the lower sideband is not needed in the modulated and linear amplifier stages, and the

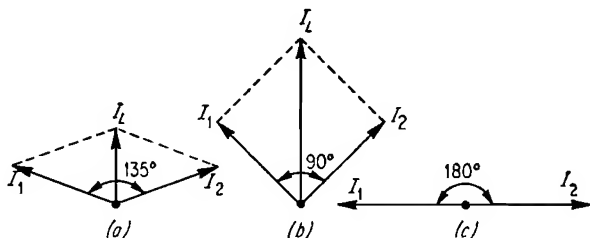


FIG. 6-21. Phase-to-amplitude modulation vector diagrams. (a) Unmodulated; (b) 100% positive; (c) 100% negative. (Courtesy of Electronics.)

passband of these amplifiers is centered on the upper sidebands. This permits the amplifiers to operate at somewhat more efficiency than if they were adjusted for the full 9-Mc bandwidth in a dsb system.

6-11. Phase-to-amplitude Modulation. Another method of obtaining amplitude modulation is the *phase-to-amplitude* or *outphasing* method (37-39). This permits the application of modulation at low power levels. The amplifiers following the modulated stage may be operated at relatively high efficiency and without any special requirements for linearity. Briefly, the method employs two identical phase-modulated transmitters fed from a common r-f generator and fed to a common load. Figure 6-21 shows vector diagrams which illustrate the operation. The unmodulated condition is shown in Fig. 6-21a, where the output currents I_1 and I_2 from the two transmitters are 135° apart in phase. The resultant current in the load is I_L . When this angle is reduced to 90° , I_L will double although I_1 and I_2 have not changed. When the angle is increased to 180° , I_L will drop to zero. Figure 6-21b depicts the 100 per cent positive condition, and c is the 100 per cent negative condition. It is seen that it is only necessary to change the phase angle of I_1 and I_2 by 22.5° to obtain 100 per cent modulation.

Since the final amplifier delivers varying amounts of power at substantially constant r-f voltage, the outphasing method is effectively a load-impedance modulation method. The load-impedance variation causes two undesirable effects on the amplifier circuits. (1) The loaded Q changes, and (2) the load has a reactive component which changes. The load variation is given by

$$R_L = 2R_L \cos^2 \frac{\theta}{2} + jR_L \sin \theta \quad (6-7)$$

In practice, the reactive component is tuned out of each amplifier at the unmodulated condition. The loaded Q is kept from reaching too high a value by adjusting the amplifiers so that the angle of flow is long enough for the tubes themselves to provide some resistance loading. These two effects do not seriously reduce the amplifier efficiency, and the method has been used successfully in both Europe and the United States for both television and standard a-m broadcast transmitters. Methods of generating phase modulation are discussed in Chap. 7.

6-12. Negative Feedback. The use of negative feedback may be effective in reducing distortion and noise and improving frequency response in a modulated transmitter. Voltage amplifiers for amplifying the input signal of all types of transmitters and the modulators for a-m transmitters will generally have internal feedback circuits. In addition, it may be desirable to rectify a portion of the transmitter output signal and feed this back into the signal input so that distortion introduced by the r-f amplifiers will be reduced. In some cases it may be advantageous to apply direct r-f feedback without first rectifying (40). The latter method is used to improve linearity in r-f amplifiers used for single side-band. In any case, application of negative feedback around several stages must be carried out with extreme care to avoid the possibility of "singing" or oscillation of the whole system. In general, the output signal is reversed in phase and fed back to the input. The over-all gain of the system will be reduced in proportion to the amount of the feedback and distortion or noise introduced by the amplifier will tend to be canceled out. The difficulty is that the phase shift of the signal which is passed through a system of amplifiers generally changes with frequency. If a frequency exists at which the feedback signal is in phase with the input signal, the feedback is no longer negative but positive, and, if the gain is sufficient, oscillation will occur. It then becomes necessary to do one of three things:

1. Reduce the amount of feedback.
2. Reduce the amplifier gain at the frequency at which positive feedback occurs.
3. Alter the phase characteristics of the amplifier or feedback path.

A complete discussion of negative feedback is a large subject, and a number of books and papers have been written on this subject (41-47). Analytic methods of determining whether or not an amplifier will be stable may be found in the literature on circuit theory and network analysis.

Distortion that is caused by overloading an amplifier cannot be reduced by the use of feedback, and, in some cases, negative feedback will make the condition worse because the overloading may introduce phase shifts that will cause the circuit to oscillate. An amplifier must not be operated in its nonlinear regions and also must have some response over the desired frequency range before feedback is applied.

With high ratios of feedback, the gain of the circuit is more dependent on the components of the feedback than on the components in the amplifier. When higher harmonic products are to be eliminated by feedback, the feedback path must pass these components. When detected radio frequency is to be used for feedback, the detector and any amplifier following the detector should be capable of passing much higher frequencies than those to be transmitted.

The amount of feedback that is being applied to an amplifier is commonly expressed in terms of the resulting reduction in amplification.

$$\text{Feedback} = 20 \log \frac{\text{gain without feedback}}{\text{gain with feedback}} \quad \text{db} \quad (6-8)$$

The noise and distortion in an amplifier will be reduced by the same ratio as the gain reduction, provided that the unwanted products are actually being generated in the amplifier and provided also that the unwanted products are not being generated by a sharply nonlinear operation such as driving the plate current to cutoff. The phase angle of the unwanted products being fed back must also be 180° out of phase with the distortion products being generated in order to obtain the optimum reduction. For example, an amplifier having 10 per cent distortion without feedback, will have only 1 per cent distortion if 20 db of feedback is applied with the correct phase relationship.

Negative Feedback Modulation. Negative feedback can be used to *remodulate* a transmitter without reducing the gain of the system by generating a difference voltage in a differential amplifier. A block diagram of this method is shown in Fig. 6-22 (51,52). If the output voltage is identical with the input voltage, no difference voltage will be generated, and no remodulation will occur. When a linear amplifier is being corrected, both the voltages being compared will be modulated r-f voltages, and these will be applied to linear detectors which will provide the inputs for the differential amplifier. If the detectors are identical, their nonlinearities will cancel each other.

Distortion in a modulated amplifier may also be corrected by applying the modulating signal and the detected r-f output to a differential amplifier (53). In this case the detector must be quite linear as there is no corresponding detector on the input side to balance out any nonlinearity.

6-13. Single Sideband. *A Comparison to Conventional Amplitude Modulation.* A single-sideband (ssb) signal utilizes only one of the two sets of sideband frequencies normally generated about the carrier by an a-m signal. Since the sidebands are symmetrical about the carrier in an a-m signal, both the lower and upper sidebands can be considered as "saying the same thing." Thus for communication purposes, either of the two sidebands can be dispensed with. This would leave a carrier signal and one sideband for communication. However, the only purpose of the carrier is to serve as a reference frequency when the signal is

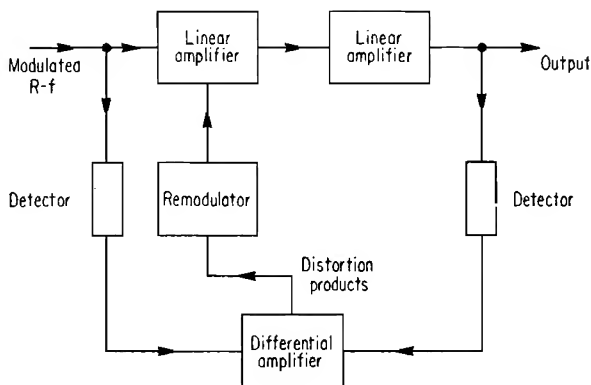


FIG. 6-22. Distortion correction by remodulation. (From *Proc. IRE.*)

received and detected at the receiver. Since the carrier signal is only used at the receiving end, one might naturally ask: Why generate this signal at the transmitter, radiate it through hundreds or thousands of miles just to use it at the receiver? Generating this signal at the receiver would save a considerable amount of power at the transmitter since the carrier power represents two-thirds of the total radiated power in a conventional, dsb, 100 per cent modulated a-m transmitter. From this we can begin to see some of the inherent advantages of the ssb system of communication. While the ssb system is useful and practical in many applications, it has enough disadvantages to make it uneconomical and cumbersome in other applications.

6-14. Advantages of Single Sideband. *Spectrum Conservation.* One of the obvious advantages of ssb modulation is its saving in spectrum space. Since the carrier and one set of sidebands are not present in the output, a single sideband will occupy less than half the spectrum space of the conventional a-m signal. This assumes that the suppressed side-

bands are sufficiently suppressed so that they will not interfere with other signals on the same frequency as the suppressed sideband. The amount of suppression required to achieve this will vary with the power output of the transmitter; however, 35 to 40 db of sideband suppression can be regarded as the minimum under normal circumstances.

Transmitter Ratings. An ssb transmitter will require a power rating of only one-eighth (9 db) the peak power rating of a conventional dsb a-m transmitter, in order to produce the same amount of sideband power. For example, if a 100 per cent modulated a-m transmitter has a carrier power of 100 watts, the power in each sideband will be 25 watts. The total sideband power will be the sum of the two sideband powers or 50 watts. The average power output of the transmitter will be 150 watts.

Since the peak voltage of a modulated r-f envelope is twice the carrier voltage, the peak power will be four times the carrier power. For the a-m transmitter of the example cited, the peak power will be 400 watts. However, an ssb transmitter need only produce 50 watts to equal the sideband power produced by a conventional a-m transmitter with a peak power of 400 watts.

In the above example, the transmitters were assumed to be modulated. However, during standby, or periods of no modulation, the ssb transmitter has the added advantage of reduced power input. This is because most ssb transmitters use class B or AB amplifiers for the final stage. In this type of amplifier, the plate current varies with the input signal. Thus, there is no power input, or at least very little, to an ssb amplifier during periods of no modulation. Since the conventional a-m transmitter must always produce the carrier power, it cannot claim this advantage.

6-15. Methods of Generating an SSB Signal. The purpose of any method of generating an ssb signal is to suppress the carrier and the unwanted sideband and to amplitude-modulate the remaining r-f sideband. There are two methods in common use that will accomplish this objective. These are the *filter method* and the *phasing method*. Each method is capable of excellent performance when properly adjusted. However the filter method, which is usually more costly initially, is generally more stable in operation and is most commonly used in commercial equipment. The phasing method finds its greatest use in amateur radio applications.

The Filter Method. The basic elements used in the filter method of generating an ssb signal are shown in Fig. 6-23. In this method, the input modulation and r-f carrier input frequency are combined in the balanced modulator. This circuit combines these two signals and provides an output signal consisting of the two sidebands of an a-m signal minus the carrier. This dsb signal is then fed through a filter, which

will pass only one of the two sidebands. Thus the output of the filter consists of one ssb signal.

This method of generating an ssb signal is often described as the *brute-force method*, since a great deal of the sideband and carrier suppression depends on the characteristics of the filter. The required suppression varies from 35 to 80 db for most ssb applications. At the same time this filter must pass virtually unattenuated frequencies only a few hundred cycles away. Quartz crystal filters with this characteristic have been constructed at frequencies from 100 kc to 5 Mc. Electromechanical filters have been commonly used at frequencies from 100 to 500 kc, while *LC* filters have been used from 25 to 100 kc.

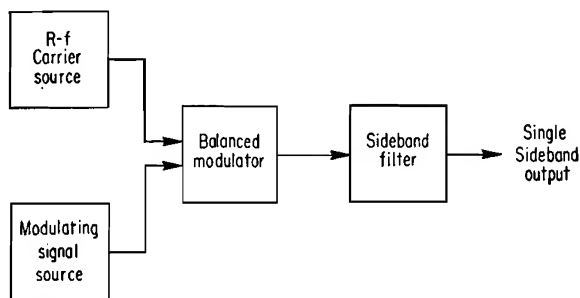


FIG. 6-23. Filter method of generating an ssb signal.

The type of filter employed in the ssb exciter also determines the output frequency of the ssb exciter. Naturally, the transmitter circuitry is simplified if the ssb-exciter output frequency is the same as the desired operating frequency. This means that no intermediate stages of frequency translation are required. However, this is not always possible. In general, the higher exciter output frequencies are desirable for transmitters intended for the vhf or uhf ranges since fewer stages of frequency translation will be required.

The Phasing Method. In the phasing method, the undesired sideband is removed by a process of phase-shifting and balanced circuits. The principle and operation is not quite as obvious as the filter method as perhaps indicated in the basic block diagram of the system shown in Fig. 6-24. The heart of this system is the wideband 90° shift network. The amount of sideband suppression depends on the constancy of the 90° of phase shift and the frequency response over the audio-modulation frequencies.

A second phase-shift network is required for the r-f signal. This phase shift is not quite as difficult to achieve, mainly because the shift is required to be 90° only at one frequency.

In the block diagram of this system, the phase-shifted r-f and audio-modulating signals are fed into one balanced modulator, and the unshifted r-f and modulating signals are fed into another balanced modulator. All these balanced modulators are identical in operation, and, as a matter of fact, they are also identical in operation with those required for use in the filter method previously described. The output of each of the balanced modulators consists of two sidebands. The carrier signal is suppressed by the balanced modulator. The outputs of the two balanced modulators are combined in a tuned circuit to form the ssb signal.

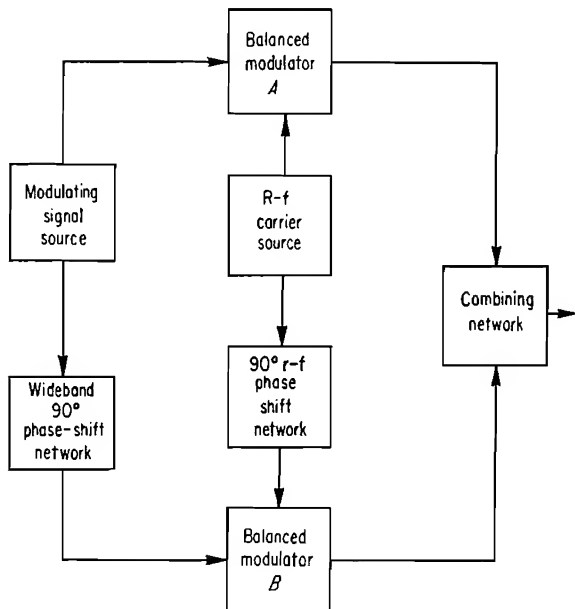


FIG. 6-24. The phasing method of generating an ssb signal.

The mechanism by which the ssb signal is generated by combining the outputs of the balanced modulators is not obvious. However, by examining the vectorial relationships that exist between the sidebands of a standard a-m signal, and then in the outputs of the balanced modulators of Fig. 6-24, the mechanism of ssb generation by this method becomes a little clearer.

The vector relationships between the sidebands and the carrier for a standard a-m signal are shown in Fig. 6-25. The r-f carrier vector is actually rotating around at whatever the carrier frequency is in cycles (or revolutions) per second. The two sideband vectors are rotating around the end of the carrier vector at a rate equal to the modulation frequency. These vectors represent respectively the upper and lower sidebands and are rotating in opposite directions. When they are in

phase, i.e., both vectors are in the same direction, a modulation peak is indicated. Similarly, when the sideband vectors are in opposite directions, i.e., the vectors cancel, a modulation dip or valley is indicated.

With this picture in mind, the vectorial relationship that exists at the output of each of the balanced modulators of Fig. 6-24 can be examined. However, in order to compare the two vector diagrams, it is necessary to fix the vectors in time to some reference, which in this case will be

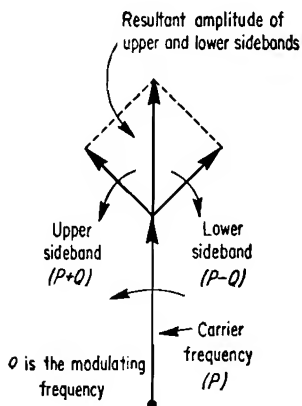


FIG. 6-25. The vector relationship for an a-m signal.

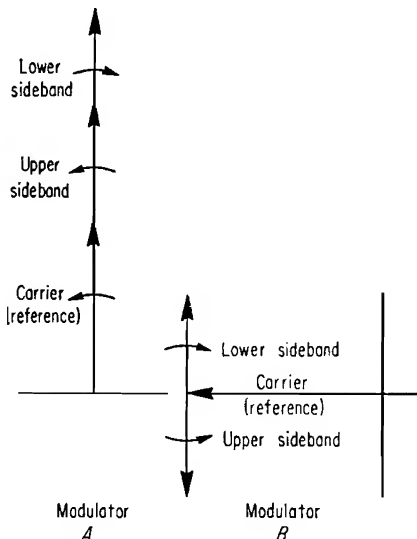


FIG. 6-26. Vector relationships of balanced modulators of Fig. 6-24.

the carrier and modulation source indicated in Fig. 6-24. The vector relationships in the upper balanced modulator *A* of Fig. 6-24 are shown in Fig. 6-26*a*. The vectors of Fig. 6-26*b* indicate the output of the lower balanced modulator *B* in which the carrier as well as the modulation sidebands are each displaced from their reference phases by 90° .

If the two vector diagrams of Fig. 6-26 are added together, as they are in the phasing method of generating an ssb signal, the result will be a carrier and upper sideband only. The lower sideband will be canceled in this case. However, since the balanced modulator eliminates the carrier in the output, the output will actually consist of only the upper sideband. By making the audio phase shift lag by 90° instead of lead by 90° , the lower sideband can be made to appear in the output instead of the upper sideband.

The actual generation of a satisfactory and usable ssb signal requires care in both the adjustment and design. For example, for the sidebands to cancel out in the manner just described, the signals must be exactly equal in amplitude and phase. In addition, these adjustments should be stable, i.e., once the controls are set, the amplitude and phase rela-

tionships should not vary. The networks through which each of the signals are passed should have the identical frequency response. Otherwise, at some frequencies the two signals will not cancel in amplitude. An additional requirement is that the two modulating signals fed into the two balanced modulators maintain the 90° phase-difference relationship over the entire band of modulating frequencies. Ideally, the sideband cancellation will then be complete, and the unwanted sideband will not appear in the output.

Although the phasing method of generating an ssb signal requires more care and adjustment during use, it makes up for these disadvantages by its versatility and relatively low initial cost. For example, the system is capable of generating a standard dsb a-m signal in addition to the ssb signal simply by disabling one of the balanced modulators and upsetting the balance of the remaining balanced modulator. The system is also capable of generating either upper or lower sideband by reversing the modulation input signal to the balanced modulators.

6-16. The Balanced Modulator. The balanced modulator plays an important part in the generation of ssb signals. In the filter method of generating an ssb signal, the balanced modulator partially relieves the filter from completely suppressing the carrier, and, in the phasing method, the balanced modulator is completely responsible for suppressing the carrier signal. Thus the feature that makes the balanced modulator desirable for ssb applications is the balanced circuitry that enables the carrier signal to be canceled out so that it does not appear in the output.

The basic purpose of the balanced modulator is the same as any other amplitude modulator or mixer circuit. A carrier signal and modulating signal are fed into the circuit and mixed in the nonlinear elements of the modulator to provide the sum and difference frequencies of the carrier and modulating frequencies. In modulators other than the balanced type, the carrier will also appear in the output. However, as previously indicated, the carrier signal is made to balance or cancel out. The sum and difference frequencies are more conventionally called the *upper* and *lower* sidebands.

The nonlinear elements in the balanced modulator may be either crystal diodes or vacuum tubes. In general, the balanced modulators using crystal diodes will maintain their balance over a longer period and will require less adjustment.

Since the balanced modulator suppresses the carrier signal by means of its balanced circuitry, the degree of carrier suppression depends on the care taken in constructing and adjusting the balanced modulator. Ordinarily, 35 to 40 db of carrier suppression can be expected with well-balanced, well-shielded circuits. There are many types of balanced modulator circuits, some of which are discussed in the following pages.

The Shunt-bridge-balanced Modulator. A typical shunt-bridge-balanced modulator is shown in Fig. 6-27. With no audio input to the balanced modulator, no radio frequency will be present in the output circuit, owing to the balance of the bridge. However, when an audio signal is impressed across the bridge, the balance of the bridge is upset by the conduction of the diodes. This enables a flow of radio frequency through the output transformer. In practice, the amplitude of the r-f signal is 10 to 20 times that of the audio-modulating signal. This is done to minimize the distortion of the output signal. For germanium and copper oxide rectifiers, the r-f voltage is usually 2 to 6 volts.

A second type of shunt-bridge-balanced modulator is shown in Fig. 6-28. This second type used only two diodes in place of the four used in

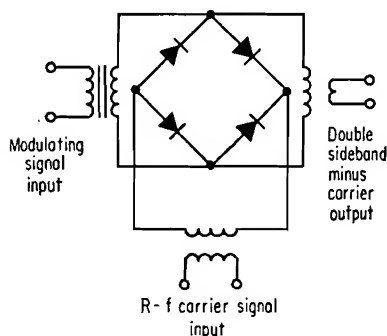


FIG. 6-27. A four-diode shunt-bridge-balanced modulator.

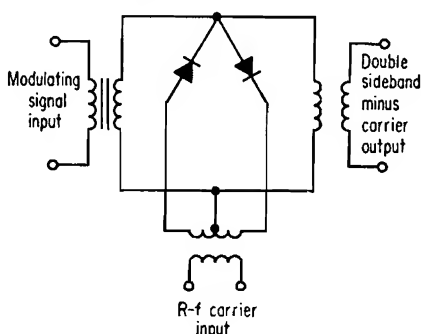


FIG. 6-28. A two-diode shunt-bridge-balanced modulator.

the previous example. The use of two diodes necessitates a tap on the r-f input transformer; however, the balance of the circuit becomes a somewhat simpler problem in a practical circuit if only two diodes are involved.

In order to obtain efficient operation in the shunt-connected configuration, the modulating source must present a low impedance to the audio-modulating frequencies and a high impedance to the r-f frequencies. The impedance of the output circuit should be a low impedance at the r-f carrier frequency and a high impedance at the modulating frequency.

Series-bridge-balanced Modulator. The basic circuit of the series-bridge-balanced modulator is shown in Figs. 6-29 and 6-30. As before, this type of balanced modulator can make use of either two or four diodes. The series modulator is sometimes favored because the impedance of the modulating and r-f source is more conveniently accommodated. In the series modulator, the impedance of the modulator must be high for the modulating frequencies and low for the r-f carrier frequency. In addition, the r-f output circuit must present a high impedance to the r-f carrier frequency and a low impedance to the modulating frequency.

Lattice or Ring-balanced Modulator. The basic circuit of the ring or lattice modulator is shown in Fig. 6-31. In this circuit, the r-f carrier will flow through the output transformer at all times, but in opposite directions. In this way, the carrier is balanced out. The mixing action

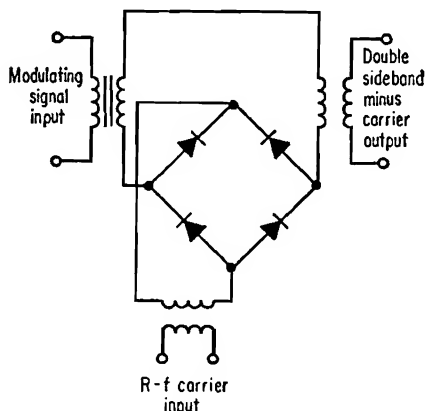


FIG. 6-29. A four-diode series-bridge-balanced modulator.

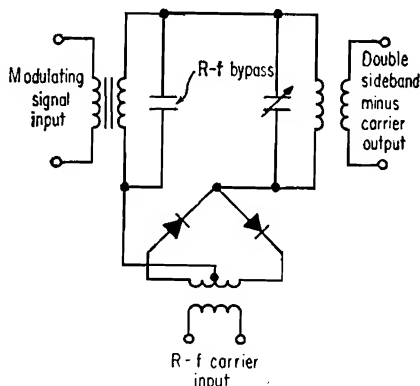


FIG. 6-30. A two-diode series-bridge-balanced modulator.

takes place between the audio-modulation signal and the r-f signal as before, and produces the sidebands that appear in the output. The input and output impedance of this type of circuit is in the order of 600 ohms. The r-f source impedance is even lower and, in practice, is usually a link from the r-f source.

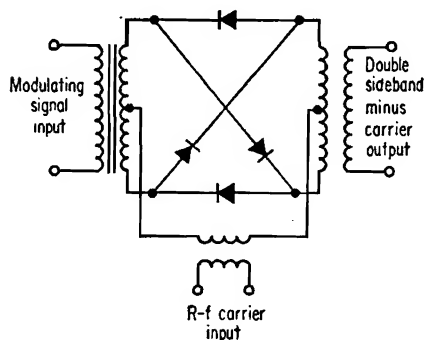


FIG. 6-31. Lattice or ring-balanced modulator.

This circuit is commonly found in telephone ssb applications and is not generally used at frequencies above 100 kc.

The Push-Pull Balanced Modulator. There are many types and kinds of balanced modulators using tubes. However many of these types are variations of two basic types. These are the push-pull and the push-push types. The basic circuit of the push-pull modulator is shown in Fig. 6-32. In this type of balanced modulator, the modulating signal is introduced

into the grid in push-pull, in much the same manner as any push-pull amplifier. Thus, instantaneously, the grids have a modulating signal applied that is of opposite polarity. The r-f signal is introduced into the grid in parallel. Thus both grids have an r-f signal that is of the same instantaneous polarity.

The two signals will mix in the tube and generate the sidebands. In the output circuit, the r-f signal will balance out in the plate circuit since the r-f signal introduced into the grid will result in a plate current that is flowing in opposite directions in the output transformer and hence will cancel out.

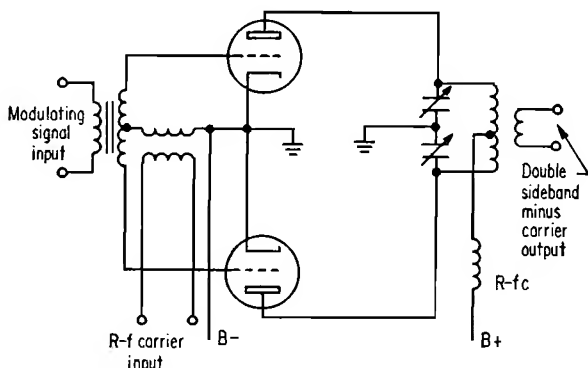


FIG. 6-32. A push-pull balanced modulator.

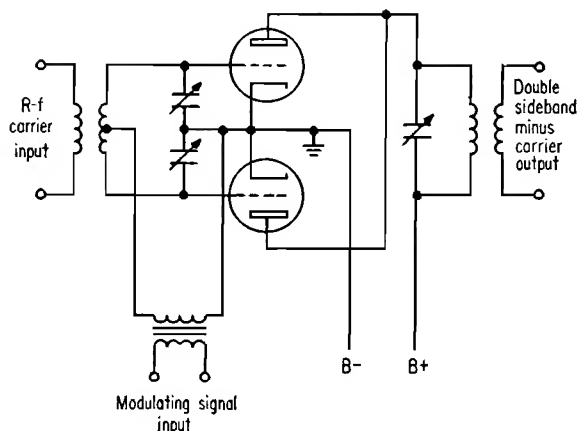


FIG. 6-33. A push-push balanced modulator.

The modulating signal does not appear in the output simply because the tuned circuit is tuned to the radio frequency and will act as a low impedance to the modulating frequency. Thus, the only output that appears in the output is the r-f sideband frequencies. This assumes that the balance of the r-f carrier signal in the output is perfect. In practice, a carrier-balancing arrangement is necessary to adjust for variations in the tube gain and imperfections in the transformer.

The Push-Push Balanced Modulator. The second basic type of balanced modulator using tubes is the push-push type (its basic circuit is shown in Fig. 6-33). This is in principle similar to the push-push fre-

quency doubler described in Chap. 3, except that a modulating signal is introduced in series with the grid return of each tube. Thus the modulating signal is placed in parallel with the grids and the r-f signal is placed on each grid in push-pull. Since the plates of the two tubes are connected in parallel, the r-f carrier signal will be balanced out in the plate circuit. The r-f signal on one grid will cause the plate current of that

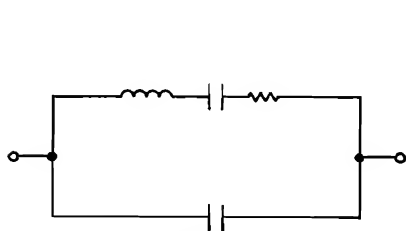


FIG. 6-34. The quartz-crystal equivalent circuit.

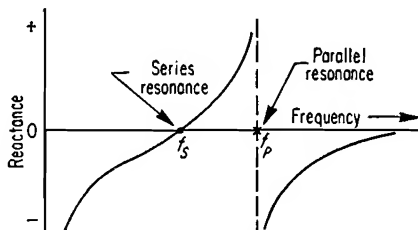


FIG. 6-35. Reactance plot of the circuit of Fig. 6-34.

tube to increase, while the grid signal on the opposite tube will cause the plate current to decrease. The r-f carrier signals will cancel each other out in the plate circuit.

The modulating signal will not appear because the output tuned circuit is a low impedance to the modulating frequency. Once again, the only frequencies that will appear in the output are the upper and lower r-f sidebands.

6-17. Quartz Crystal Filters. One of the most satisfactory methods of achieving the narrow bandwidth required in ssb transmitters is through the use of quartz crystal filters.

This type of filter, as its name implies, uses quartz crystals as the filter elements.

A quartz crystal has the electrical equivalent circuit shown in Fig. 6-34. If a reactance plot were taken of this network, the curve of Fig. 6-35 would be obtained. From this curve we can observe that a quartz crystal has two resonant frequencies, i.e., one parallel resonance and one series resonance.

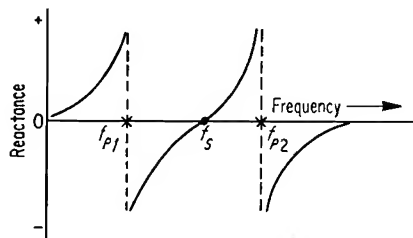


FIG. 6-36. The reactance plot obtained by shunting a quartz crystal with an external coil.

Normally these frequencies are only a few hundred cycles apart.

If a quartz crystal is shunted with an external coil, the reactance plot of such a circuit would indicate the presence of a second parallel resonance. This is shown in Fig. 6-36. This effect is known as *spreading*. This spreading effect is used in most filters to obtain the desired response. For example, in the filter shown in Fig. 6-38, the spreading is obtained

by the input and output coils. Mathematically it could be shown that the coils are effectively shunted across the crystals. The reactance plot for the crystals in the series and shunt arms of the filter are shown in Fig. 6-37. The crystals in the series arm are within 50 cps of each other.

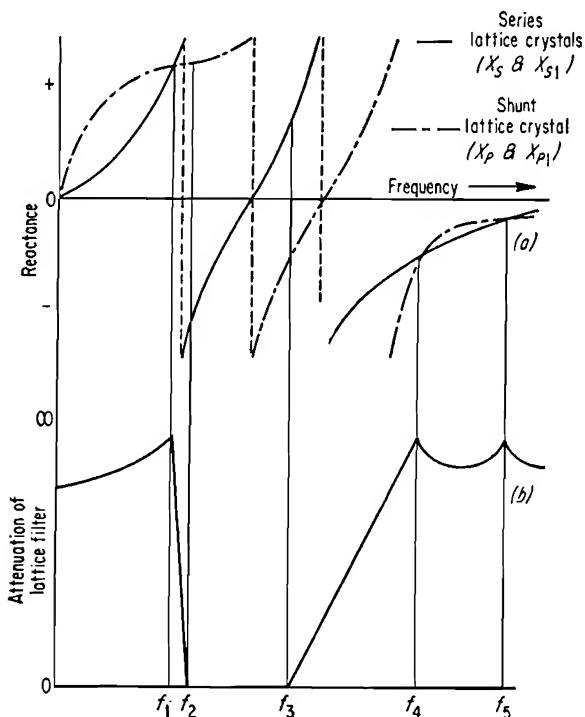


FIG. 6-37. (a) The reactance plot for the shunt and series quartz crystals for the filter in Fig. 6-38. (b) Attenuation plot.

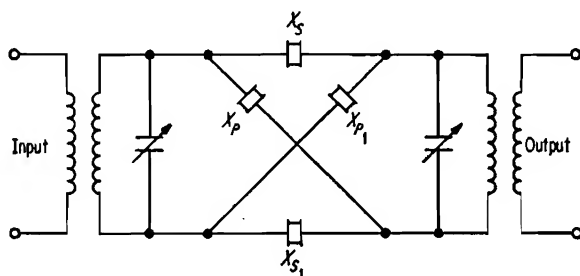


FIG. 6-38. A typical lattice-quartz-crystal filter.

Similarly the shunt crystal frequencies are within 50 cps of each other. For the plot shown, the series crystals are a few kilocycles lower in frequency than the shunt crystals.

The bandpass filtering action of the crystal filter can now be visualized

—at least qualitatively. But first it is necessary to consider the crystal filter network. The network shown in Fig. 6-38 is known as the *crystal lattice*. This network or some variation of it is actually the basis for just about every commercial crystal filter. A closer look at the configuration of this filter reveals that it is basically a bridge, and it can be redrawn to the circuit shown in Fig. 6-39. This is not the common way of drawing the crystal lattice, however, it will make the explanation of the operation somewhat clearer. When any bridge is balanced, the attenuation

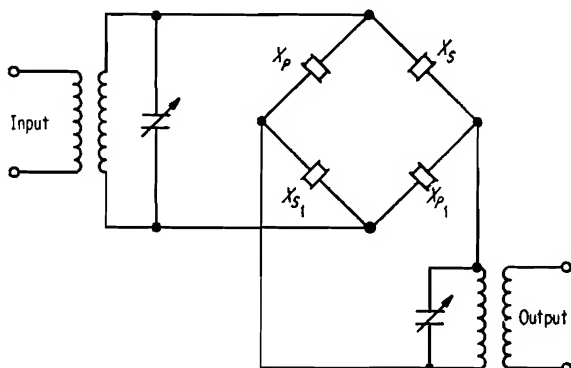


FIG. 6-39. The lattice filter of Fig. 6-38 redrawn as a bridge.

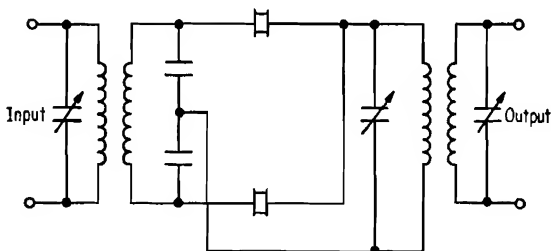


FIG. 6-40. A typical half-lattice crystal filter.

between input and output will be at a minimum. Conversely, when the bridge is not balanced, transmission will occur between input and output. Now, since the arms of the bridge are quartz crystals, the bridge will be balanced when the crystals have the same reactance value and sign; i.e., all the arms are of the same value inductance or the same value capacitance. At this point, the maximum attenuation will occur between input and output. This is indicated on the reactance plot of the shunt and series crystals used in the filter and plotted in Fig. 6-37 as f_1 , f_4 , and f_5 .

The transmission through the filter will be at a maximum whenever the reactances are equal but opposite in sign. This will occur in the bandpass region shown in Fig. 6-37*b* as f_2 and f_3 .

One of the more common variations of the full crystal-lattice filter just described is the half lattice shown in Fig. 6-40. The half crystal-lattice filter has only two crystals and has essentially the same response as the full lattice. The response of this filter will be approximately 1.2 times the frequency difference of the two crystals used in the filter.

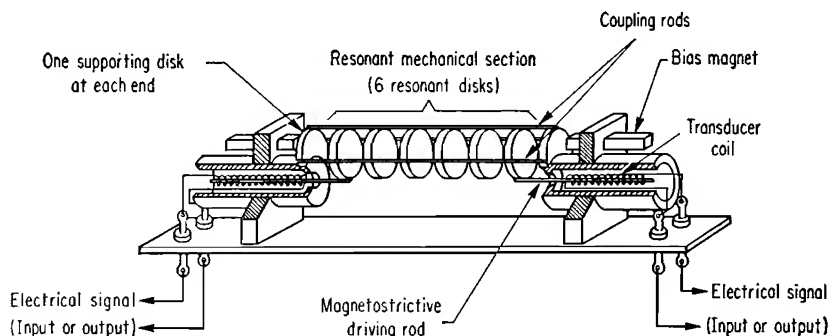


FIG. 6-41. A typical magnetostriction filter. (Courtesy of Collins Radio.)

To obtain the desired amount of suppression of the unwanted sidebands and carrier, the filters can be cascaded, i.e., connected in series.

6-18. Mechanical Filters. Another means of achieving the selectivity required in the filter method of generating an ssb signal is through the use of an electromechanical bandpass filter. A filter of this type is shown in its basic form in Fig. 6-41. This filter consists of an input transducer, a resonant mechanical section, an output transducer, and a set of bias magnets for the transducers.

The input transducer mechanically vibrates the disks in the mechanical section. This is done by magnetostriction; that is, the driving rod actually undergoes a change in length when subjected to a magnetic field. This magnetic field is the result of the driving signal current through the coil of the transducer.

The effect of magnetostriction of a nickel rod is shown in Fig. 6-42. Since the input signal is continuously alternating its polarity, a permanent magnet bias is required to place the operating point somewhere in the middle of the curve of Fig. 6-42.

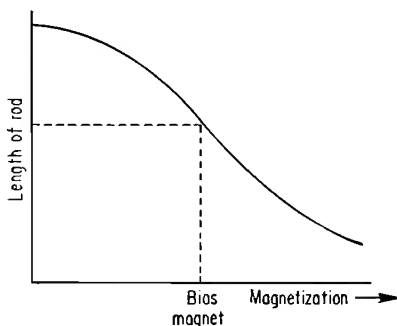


FIG. 6-42. The magnetostrictive effect of a nickel rod.

At the mechanical resonant frequency of the disks, the amplitude of the vibration increases appreciably. The resulting motion is transmitted

through the coupling rods in the filter to the output transducer. The mechanical motion of the coupling rod in the output transducer coil results in an induced voltage in the output coil.

Generally, the coils of the input and output transducers are resonated to the operating frequency of the filter. This assists in suppressing spurious resonant modes that lie outside the immediate bandpass region, as well as providing a greater driving current for the magnetostrictive driving rods.

The Q obtained from a mechanical resonator is appreciably better than a tuned circuit in the frequency range of 60 to 600 kc. In general, the Q of the mechanical filter elements is over 5,000.

6-19. Audio Phase-shifting Networks. The phasing method of generating an ssb signal requires accurate phase-shift networks for both the r-f carrier signals and for the audio-modulating signal. The audio phase-shift network is usually the most difficult to design since it must provide a constant 90° phase shift over the entire range of the modulating frequencies. In addition, the network must have a constant output voltage (for a constant input voltage) over the range of the modulating frequencies. If the phase shift or amplitude response varies, then some amount of unwanted sideband power will appear in the output. The amount of sideband suppression expressed in decibels can be determined from the following equation.

$$\text{Sideband suppression} = 20 \log \left(\cot \frac{\phi}{2} \right) \quad \text{db} \quad (6-10)$$

where ϕ is the total phase-angle error in the network.

Thus, if an angle of 1° phase-shift error is substituted in the above equation, we find that the maximum sideband suppression possible in the system is 40 db. The substitution of 2° results in a maximum sideband suppression of 35 db. This assumes that the other factors in the phasing system are theoretically perfect; i.e., the amplitude of the signals are perfectly balanced, and the phase shift of the r-f signals is exactly 90° .

The theoretical maximum sideband suppression that can result in the presence of amplitude variations out of the phase-shifting network is expressed in the following equation.

$$\text{Sideband suppression} = 20 \log \frac{200 + A}{A} \quad \text{db} \quad (6-11)$$

where A is the difference of the amplitude of the two voltages expressed as a per cent.

Thus a difference in the two modulating voltages of only 1 per cent will result in a maximum sideband suppression of 45 db; a 2 per cent

difference in the modulating voltage would result in a maximum sideband suppression of 40 db.

Audio phase-shift networks are either of the active or passive types. An active network is one in which amplifiers are used. A passive network employs no means of amplification, nor has any other source of operating power other than the input signal. Active phase-shift networks were used in earlier ssb exciters using the phasing method. However, subsequent development has produced satisfactory constant-amplitude phase-shifting networks using only resistances and capacitances. These networks can be designed to produce an almost constant phase shift over the audio speech range of 150 to 2,700 cps. The error over this range is usually no more than 2° . It is necessary to restrict the audio

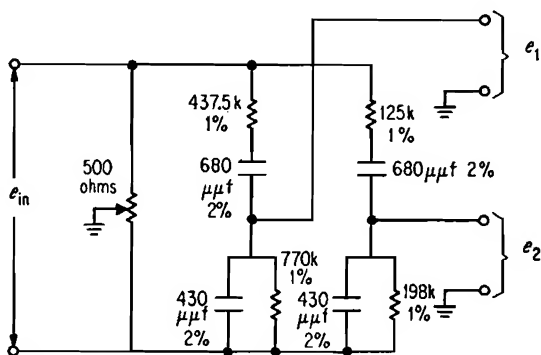


FIG. 6-43. A typical 90° passive audio phase-shifting network.

input frequencies to this band of frequency by means of a bandpass filter. If frequencies outside of the range of the phase-shifting network were allowed to modulate the transmitter, excessive amounts of unwanted sideband power would appear in the transmitter output. A typical passive audio phase-shifting network for the audio speech range is shown in Fig. 6-43.

6-20. R-F Phase-shift Networks. The phasing method of generating an ssb signal also requires that the r-f carrier signal be accurately shifted 90° . This is easily achieved if the exciter design does not require constant readjustment of the r-f phase-shifting network each time the output frequency is changed. Some typical networks used in this application are shown in Fig. 6-44. The *RLC* network shown in Fig. 6-44a has a phase characteristic that does not vary with frequency. However, the amplitude balance of the two output signals e_1 and e_2 will vary with frequency. The effect of the amplitude balance on the carrier suppression is described in Sec. 6-20.

The *RC* network shown in Fig. 6-44b also exhibits an output phase

characteristic independent of frequency. However, the amplitude balance and the input impedance will both vary with frequency.

The π network shown in Fig. 6-44c is not too widely used since the network has both a variable phase and amplitude characteristic. Thus

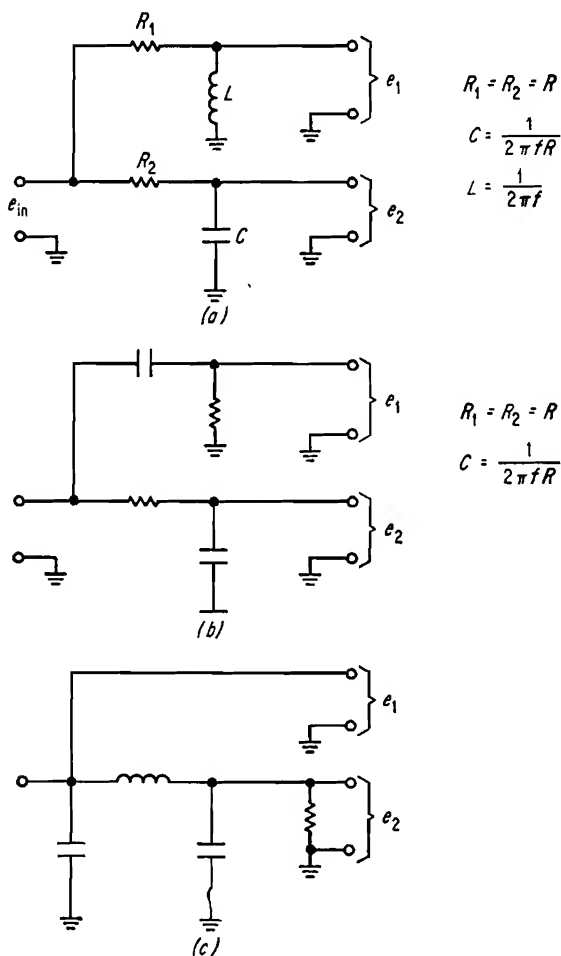


FIG. 6-44. Typical 90° r-f phase-shifting networks. (a) RLC network where f is in cycles, R in ohms, L in henrys, and C in farads; (b) RC network; (c) π network.

the performance of this network degrades rather rapidly with frequency changes. This can be calculated from Eqs. (6-10) and (6-11).

6-21. Frequency Translation in SSB Transmitters. Most ssb exciters generate the signal at a much lower frequency than the final output frequency of the transmitter. This is done mainly for convenience in the exciter design. However, to obtain the final transmitter output fre-

quency, the process of heterodyning must be used. Heterodyning the exciter signal to the final frequency preserves the original modulating-frequency relationships to the carrier. Even though the carrier is not present in the transmitter output, it will be reinserted in the receiver, so this relationship must be preserved. If the ssb exciter frequency were changed by frequency multipliers (see Chap. 3), this relationship would

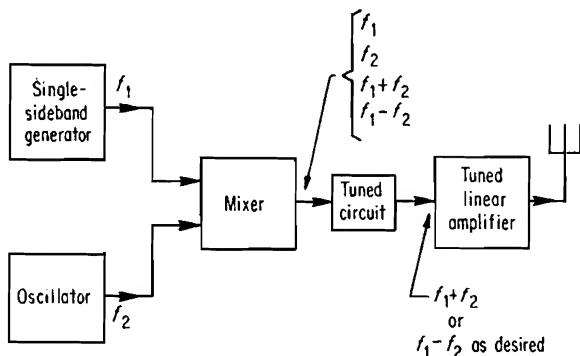


FIG. 6-45. Frequency translation used in ssb applications.

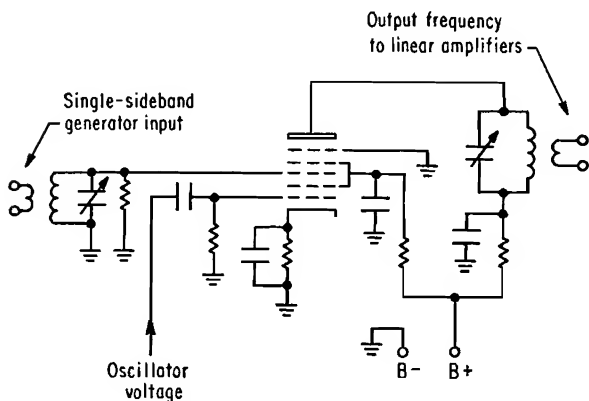


FIG. 6-46. A basic frequency mixer.

be destroyed. For example, each time the frequencies were doubled, the frequency difference between the carrier and the sidebands would also double.

The process of frequency changing by heterodyning in ssb equipment is shown in Fig. 6-45. It is actually very similar in operation to that employed in the front end of every superheterodyne receiver to produce the intermediate frequency. The mixer circuit employed in this application may be similar to the balanced-modulator circuits previously described or to the conventional mixer circuit illustrated in Fig. 6-46.

This latter circuit does not eliminate the carrier frequency; however, the tuned circuits following the mixer stage usually serve to suppress the undesired mixer-output frequencies.

If the tuned circuits do not have sufficient selectivity to reject the undesired output frequencies, the output frequency may be achieved in two (or more) steps of heterodyning. This is shown in the example of Fig. 6-47.

The selection of the difference frequency in the output of the mixer will invert the original sideband selection made in the exciter. For example, if the upper sideband was selected in the exciter, choosing the

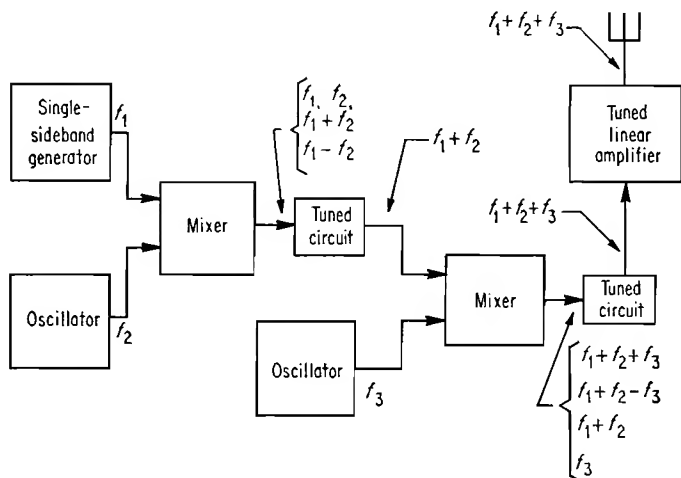


FIG. 6-47. Two steps of frequency translation to minimize undesired frequencies in the output.

difference frequency in the heterodyne frequency changer will result in the upper sideband being inverted and transmitted as the lower sideband. This is actually no problem and not detrimental to the quality of the transmission if it is recognized in the tuning and operation of the equipment. If the sum frequency is selected in the mixer output, the original sideband selection is retained in the output.

6-22. Linearity Measurement in SSB Transmitters. Normally, a ssb signal is generated at a low power level, and the final power-output level is achieved through the use of successive stages of linear amplifiers. Since one of the features of ssb transmission is the frequency-spectrum conservation resulting from the elimination of one of the sidebands and the carrier, the linear amplifiers following the ssb generator must have a high degree of linearity to prevent reintroducing additional odd-order distortion-product frequencies in the radiated output. However, even

though a linear amplifier is designed in accordance with the principles outlined in Sec. 3-16, additional adjustment of the operating parameters is generally required to achieve the desired results.

Linear amplifiers intended for ssb applications are generally adjusted for optimum linearity by means of a *two-tone* test. This test requires passing two r-f signals through the linear amplifier. These signals are usually separated by approximately 1,000 cps. The frequency difference is not particularly critical in the test—as long as the frequencies are within the normal passband of the amplifier under test.

When two r-f signals of equal amplitude are added together, as they would be in the two-tone test, the result would be a waveform similar to that shown in Fig. 6-48. This provides a convenient method for obtaining an r-f signal with an amplitude variation from zero up to the sum of the peaks of the two input signals.

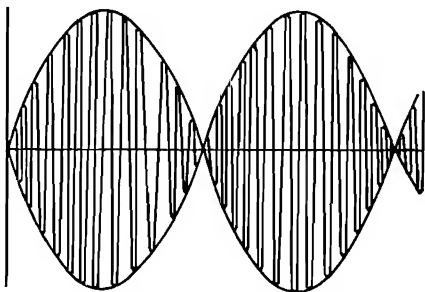


FIG. 6-48. R-f envelope of two equal-amplitude r-f signals.

If the output signal is compared to the input signal on an oscilloscope, the result can be an accurate representation of the linearity of the amplifier. The exact connection of the oscilloscope will vary according to the desired pattern and the type of exciter. These will be discussed in detail.

Generating a Two-tone Test Signal. There are a number of methods of generating the two r-f signals for the two-tone test. The choice of the method depends on whether the purpose of the test is to check the entire ssb transmitter or just the linear amplifier.

One method, and the simplest, which can be used for checking linear-amplifier performance is to adjust the carrier-insertion control or carrier-balance control on the ssb exciter to permit some carrier output. If a single tone is applied to the ssb exciter input, then the r-f output of the exciter will consist of a carrier plus one sideband signal. If the carrier level is adjusted to be equal to the sideband signal, the result will be an output signal similar to that shown in Fig. 6-48.

If it is desired to check the linearity of the entire ssb transmitter, including the exciter, then it is necessary to use two separate audio generators, as shown in Fig. 6-49. Once again, the levels of the two input signals are adjusted to produce two equal amplitude sidebands.

Monitoring the Two-tone Test Signal. The two-tone test signal can be monitored by the method previously described in Sec. 3-16. In addition, the methods shown in Fig. 6-49 and Fig. 6-52 can also be used.

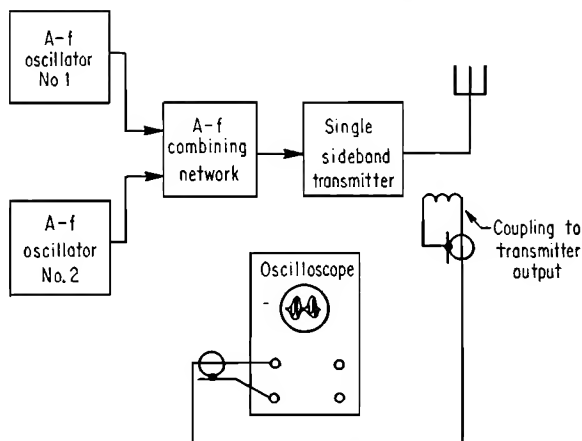
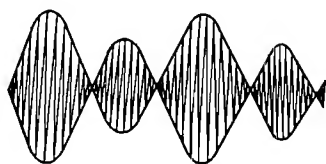


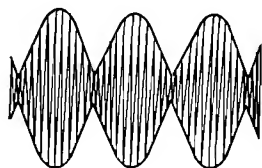
FIG. 6-49. Test set up for generating a two-tone test signal.

The typical oscilloscope patterns that result from the method shown in Fig. 6-49 are shown in Figs. 6-50 and 6-51.

The monitoring method shown in Fig. 6-52 is applicable to only the phasing type of ssb transmitters. In this method, the input signal to one of the balanced modulators is disabled. This will result in a dsb signal without a carrier. When an oscilloscope is connected as shown in Fig. 6-52, the resulting

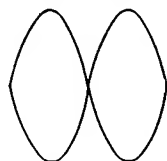


(a)

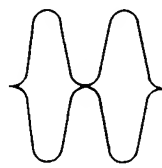


(b)

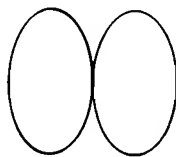
FIG. 6-50. Improper test waveforms. (a) R-f carrier leakage through modulator; (b) unequal r-f signals.



(a)



(b)



(c)

FIG. 6-51. Two-tone test waveforms. (a) Ideal; (b) grid bias too high; (c) excessive r-f drive or incorrect loading.

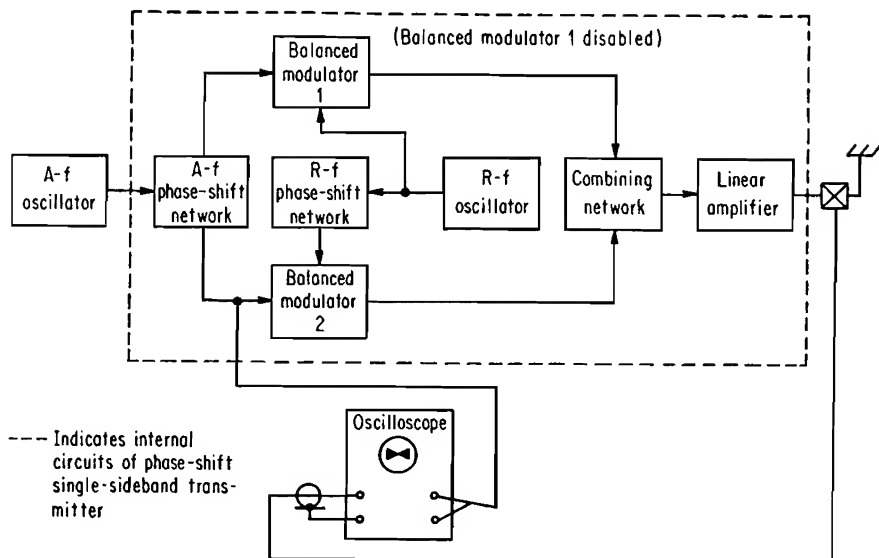


FIG. 6-52. Test setup for generating and monitoring a two-tone test signal in the phasing-type ssb transmitter.

oscilloscope pattern will be a trapezoid. Some typical trapezoid patterns are shown in Fig. 6-53.

A fourth method of observing the results of the two-tone test is through the use of a spectrum analyzer. This is an instrument which will display, on an oscilloscope, all the sidebands that may result from any existing non-linearities within the amplifier. These sidebands or odd-order distortion products can be observed and measured to determine amplifier linearity.

6-23. Signal-to-distortion Ratio.

The linearity of an amplifier can be measured in terms of the distortion generated by the amplifier as the signal-to-distortion ratio. This can be done by means of the spectrum analyzer previously mentioned, in which the amplitude of the odd distortion product can be measured. Another method is to apply two tones to the ssb exciter input. The

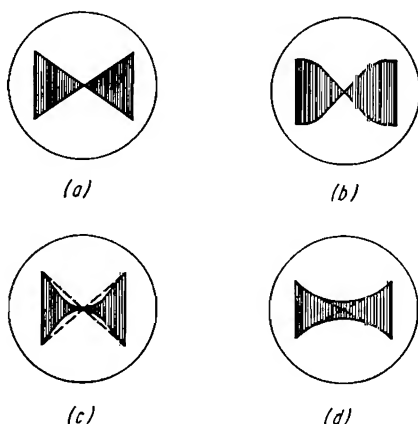


FIG. 6-53. Double-trapezoid ssb waveforms. (a) Ideal; (b) excessive r-f drive or incorrect loading; (c) excessive d-c grid bias; (d) unequal r-f signals resulting from leakage through disabled modulator or unequal sidebands.

amplitude of the demodulated output as obtained from an ssb receiver is compared to the amplitude of the distortion products obtained by passing the demodulated output through a band-pass filter. The resulting ratio is the signal-to-distortion ratio. This ratio is usually expressed in decibels.

6-24. Sideband Suppression. The amount of sideband suppression in an ssb transmitter can be determined by observing the r-f output envelope on an oscilloscope. If a single tone is fed into the exciter input, the r-f output envelope theoretically should appear as shown in Fig. 6-54a. However, if output consists of more than one frequency, the r-f envelope will take on the appearance of a partially amplitude-modulated r-f

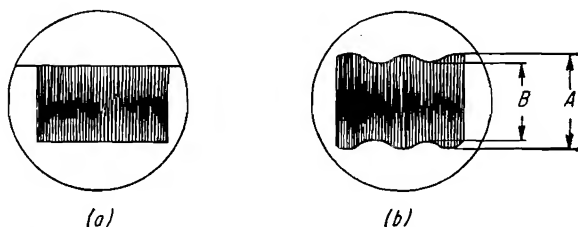


FIG. 6-54. (a) R-f envelope for a single tone fed into ssb exciter. (b) R-f envelope with incomplete sideband suppression.

envelope. This is illustrated in Fig. 6-54b. If the crest and valley of the resulting a-m envelope is measured, and the values substituted in the following formula, the amount of sideband suppression can be estimated.

$$\text{Sideband suppression} = 20 \log \frac{A + B}{A - B} \quad \text{db} \quad (6-12)$$

A second method of measuring the sideband as well as carrier suppression is through the use of a spectrum analyzer. This enables measuring the amplitude of both the desired sideband and the undesired sideband or the carrier. The ratio of these factors to the desired sideband expressed in decibels, will give the amount of suppression.

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58. Roberts, W. V. B.: Magnetostriction Devices and Mechanical Filters for Radio Frequencies, *QST*, part I, p. 24, June, 1953; part II, p. 28, July, 1953; part III, p. 32, August, 1953.
59. Kosowsky, D. I.: High Frequency Crystal Filter Design Techniques and Applications, *Proc. IRE*, p. 419, February, 1958.

ANGLE AND PULSE MODULATION

7-1. Angle Modulation. In angular or angle modulation, the phase angle of a sinusoidal carrier is varied by the modulating wave. The two practical forms of angle modulation are *phase* and *frequency* modulation (1-4). The signal-to-noise ratio after detection at the receiver will be improved over the carrier-to-noise ratio of the signal entering the receiver by an amount which depends on the amount of phase modulation. This was discussed in Chap. 1 and shown in Fig. 1-1.

If the modulating voltage applied to a phase modulator has the form $\cos 2\pi f_m t$, the expression for the resulting phase-modulated wave will be

$$e = E_c \cos (2\pi f_c t + \Delta\theta \cos 2\pi f_m t) \quad (7-1)$$

where e = voltage at any time t

E_c = maximum carrier voltage

f_c = carrier frequency

f_m = modulating frequency

$\Delta\theta$ = maximum phase deviation

Now, if the modulating voltage applied to a frequency modulator has the form $-\sin 2\pi f_m t$, the expression for the f-m wave will be exactly the same as Eq. (7-1) except that $\Delta\theta$ will be replaced by $\Delta f_c/f_m$ where Δf_c is the maximum frequency deviation.

If a frequency modulator and a phase modulator are set up and a single tone is applied so that the frequency deviation is the same, an observer with an f-m receiver will be unable to distinguish between the two. If, however, both are operated at the same time, driving transmitters on different radio frequencies, but with the same modulating tone applied in phase, and if the outputs from two identical f-m receivers are connected to a double-beam oscilloscope, it will be found that there will be a 90° difference between the audio tones obtained from the f-m and p-m transmitter. A second difference will be observed if the modulating frequency is doubled. It will be found that the amplitude of the recovered audio from the f-m transmitter will be unchanged while that from the transmitter containing the phase modulator will be doubled. A third difference will be observed if d-c is applied to the modulators and the

frequency is read. Unless the f-m transmitter has a-f-c, its frequency will change. The p-m frequency will only change momentarily.

The *phase deviation* of an angle-modulated transmitter is the maximum departure of the phase angle from what it would have been in the absence of modulation. The *frequency deviation* is the maximum departure of the transmitted frequency from the unmodulated frequency and is *one-half* of the total *frequency excursion*.

If the phase deviation is expressed in radians per second, the number obtained has been called the *modulation index*.

$$m = \frac{\Delta f_c}{f_m} = \Delta \theta \quad (7-2)$$

where m = modulation index

Δf_c = frequency deviation, cps

f_m = modulating frequency, cps

$\Delta \theta$ = phase deviation, radians per sec (1 radian = 57.3°)

For a constant amplitude of modulating voltage and changing modulating frequency, the value of m remains constant in *phase* modulators while the value of Δf_c remains constant in *frequency* modulators.

7-2. Sidebands Produced by Angular Modulation. Angular modulation is characterized by the fact that the amplitude of the r-f output voltage remains constant while the number and amplitude of the side frequencies or upper and lower sidebands is determined by the degree of modulation. Energy is removed from the carrier and transferred to the sidebands, and for certain values of modulation the carrier amplitude

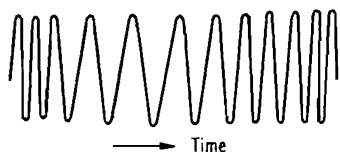


FIG. 7-1. F-m wave plotted on a linear time base.

will be reduced to zero. Increasing the modulation index beyond this point will start to increase the carrier amplitude again. Examination of a frequency-modulated signal in the *time domain* which means examining it with an ordinary oscilloscope using a linear time sweep, will yield practically no information.

Much more can be learned by observing the signal in the *frequency domain* or observing the side frequencies with a spectrum analyzer. The amplitudes of the sidebands are of interest for two principal reasons: (1) determination of required bandwidth in transmitter amplifier stages, and (2) determination of exact modulation index. Figure 7-1 shows an f-m signal plotted against a linear time base. Figure 7-2a shows the sideband distribution for a deviation of 75 kc and a modulating frequency of 7.5 kc. Figure 7-2b shows it for a modulating frequency of 3.75 kc. The *significant* sidebands are usually considered to be those that exceed 1 per cent of the unmodulated carrier. The

significant sidebands extend beyond the frequency-deviation limits by a greater amount as the modulating frequency is increased. For distortionless transmission, f-m or p-m signals will have symmetrical amplitudes on either side of the carrier if the modulating signal is an undistorted sine wave.

Conservation of bandwidth by suppression of upper or lower sidebands in f-m signals is not usually attempted. If any sidebands are suppressed, the output voltage will no longer remain constant, and the result will be a composite amplitude- and angular-modulated signal. Conservation of the spectrum is better achieved by reducing the deviation. An unsymmetrical sideband distribution will also occur when simultaneous amplitude and angular modulation are used. Such systems have been developed (5).

7-3. Sideband Characteristics.

The expression for an f-m or p-m signal may be written as an infinite series which has coefficients which

are known as *Bessel functions*. Curves of the Bessel functions for the carrier and sidebands for various amounts of modulation are given in a number of places in the literature (6-10). Values of particular interest are the points at which the carrier and sidebands pass through zero. If the f-m signal is observed on a spectrum analyzer, the amplitude of the modulating signal can be varied until the carrier or one of the pairs of sidebands is observed to go through zero. The frequency deviation can then be determined from Eq. (7-2). A few values of m at which disappearance occurs are given in Table 7-1. For example, suppose

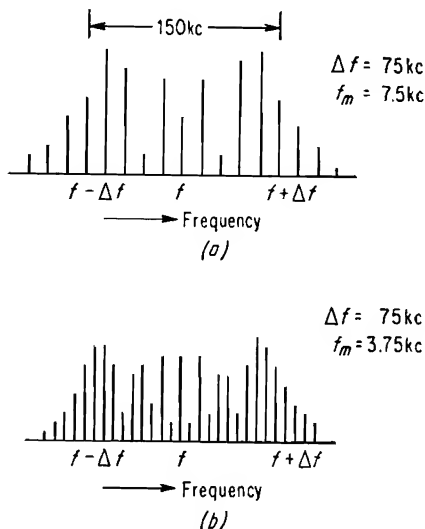


FIG. 7-2. Sidebands of f-m waves plotted on a linear frequency base. (a) Modulating frequency 7.5 kc and frequency deviation 75 kc; (b) modulating frequency 3.75 kc and frequency deviation 75 kc.

TABLE 7-1. DEVIATION RATIOS FOR CARRIER DISAPPEARANCE

Disappearance	Carrier	First sideband	Second sideband	Third sideband
First.....	2.40	3.83	5.13	6.38
Second.....	5.52	7.02	8.42	9.76
Third.....	8.65	10.17	11.62	13.01

that an f-m transmitter is modulated with a frequency of 13,580 cps. As the audio level is increased, the carrier will drop to zero for the first time when $m = 2.40$ or the deviation is 32.6 kc. As the modulation level is further increased, the carrier will reappear, rise to a maximum value which will be about 40 per cent of the unmodulated carrier at ± 51.6 kc deviation, and then fall to zero for the second disappearance at $m = 5.52$ or $\Delta f = \pm 75$ kc.

The bandwidth required for transmission of signals having a modulation index of less than 0.5 is equal to twice the modulating frequency. For higher values of m , the bandwidth required to transmit sidebands that are 1 per cent or more of the unmodulated carrier are given in Table 7-2. For example, suppose that the frequency deviation is ± 75 kc

TABLE 7-2. BANDWIDTHS REQUIRED FOR TRANSMISSION OF SIDEBANDS WHICH ARE MORE THAN 1% OF THE UNMODULATED CARRIER

Modulation index m	To obtain bandwidth	
	Multiply modulating frequency f_m by	Multiply frequency deviation Δf by
0.5	4	8
1.0	6	6
2.0	8	4
3.0	12	4
5.0	16	3.1
10	28	2.8
15	38	2.5
20	50	2.5
25	60	2.4

and the modulating frequency is 15,000 cps. The modulation index from Eq. (7-2) is 5 and the bandwidth for transmission of all significant sidebands will be $16f_m$ or $16 \times 15 = 240$ kc. When the modulating frequency is dropped to 3,750 cps, the index becomes 20, and the bandwidth is $50f_m$ or $50 \times 3.75 = 187.5$ kc. As the modulating frequency is further reduced, the bandwidth required will come closer to twice the frequency deviation.

7-4. P-M to F-M Conversion. A p-m and an f-m wave are identical and have identical sidebands for any given audio frequency. Angle-modulation generators can be separated into two basic categories: those which generate p-m and those which generate f-m signals. It is therefore frequently desirable to employ p-m to f-m conversion circuits (11). Actually it is seldom that a transmission can be defined as purely f-m

or p-m. The broadcast band from 88 to 108 Mc is generally referred to as f-m; but above about 1 kc these transmitters swing over to phase modulation if the effect of a constant audio input signal is considered. On the other hand, police, and taxi mobile transmitters in the 152- to 162-Mc band are generally considered to use phase modulators; but again at about 1 kc and above these transmitters swing over to frequency

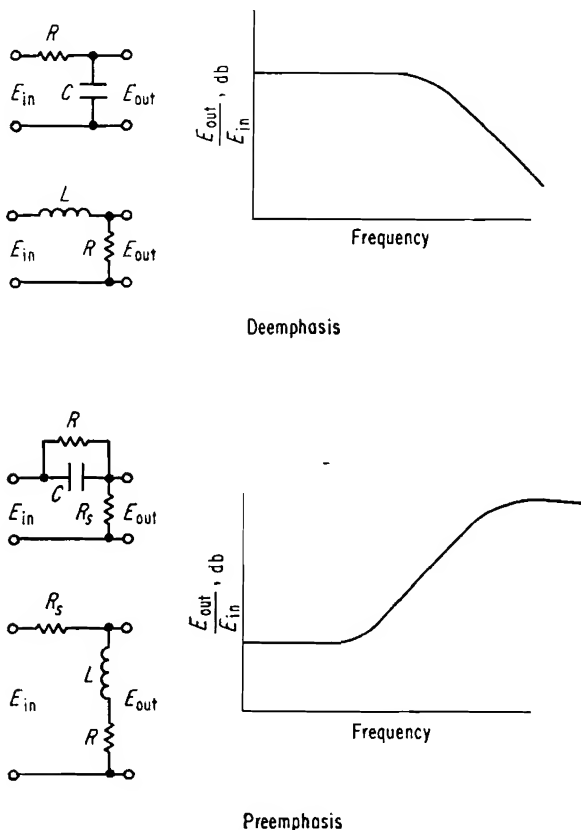


FIG. 7-3. Deemphasis and preemphasis circuits with resulting responses.

modulation. The p-m to f-m conversion is performed on the audio signal before it reaches the modulator.

In the case of broadcast transmitters, the process is generally referred to as *preemphasis* and is carried out so that corresponding *deemphasis* at the receiver will also improve the signal-to-noise ratio. In mobile transmitters the frequency response from the f-m point of view rises at a rate of 6 db per octave to about 1,000 cps, then levels off to be flat up to about 3,000 cps, and then falls off rapidly. This is done to limit the

maximum frequency deviation to 5 kc because of available bandwidth limitations in the communication bands.

Frequency modulation is converted to phase modulation by preemphasis networks, and phase modulation is changed to frequency modulation by deemphasis networks. Four combinations of R , L , and C for such networks are shown in Fig. 7-3. The frequency response of these networks rises or falls at a rate of 6 db per octave (amplitude doubles or halves as frequency is doubled) as long as the resistance is large compared with the inductive or capacitive reactance and as long as R_s is large compared with R . The rise or fall compared with a low frequency is given by

$$\text{Decibels} = 10 \log [1 + (2fT)^2 \times 10^{-5}] \quad (7-3)$$

where f = frequency, kc

T = time constant, μsec

$$= CR \mu\text{sec}$$

$$= \frac{L}{R} \mu\text{sec}$$

where C is in microfarads, L is in microhenrys, and R is in ohms.

In the case of the preemphasis networks the output-to-input voltage ratio at low frequencies will be $R/(R + R_s)$ for the LR circuit and $R_s/(R + R_s)$ for the CR circuit. For example, suppose that an RC combination is used with $R = 75,000$ ohms and $C = 0.001 \mu\text{f}$. For these values of R and C the value of T will be $75 \mu\text{sec}$. Then the rise or fall from the low-frequency value at 15,000 cps, for example, will be

$$\begin{aligned} 10 \log [1 + (2 \times 15 \times 75)^2 \times 10^{-5}] &= 10 \log (1 + 50.625) \\ &= 10 \times 1.712 \\ &= 17.12 \text{ db} \end{aligned}$$

This will be a rise from the low-frequency value for the preemphasis network and a fall for the deemphasis network.

7-5. Methods of Generating Phase Modulation. In the phase modulator, the signal is first generated in a constant-frequency oscillator, and then it is passed through some sort of device which introduces periodic phase variations. The oscillator may therefore be crystal-controlled, and this is a basic advantage of the phase modulator over a frequency modulator. The average frequency generated by a direct f-m oscillator (with the exception of directly modulated crystal oscillators) must be stabilized by some auxiliary frequency-correcting device. See Sec. 2-14.

An elementary method of phase modulation is shown in Fig. 7-4 (12). Tubes A and B have their plates connected in parallel to the output circuit but have their grids connected across a phase-shifting network. If the reactance C is chosen to be equal to the resistance R at the carrier frequency, the voltages applied to the two grids will be equal but 90°

apart in phase. Referring to the vector diagram, the output voltage OP will be the vector sum of the voltages appearing at the plates of the two tubes. Now it can be seen that, if the bias on the other grid of tube B is increased and the bias on the other grid of tube A is decreased, OA will become longer, and OB will become shorter. The new resultant OP_1 will be at an angle θ from OP . If an audio voltage is applied to these two grids in phase opposition, the output voltage will be phase-modulated in accordance with the audio modulation. This type of modulator can never produce more than $\pm 45^\circ$, and much smaller angles

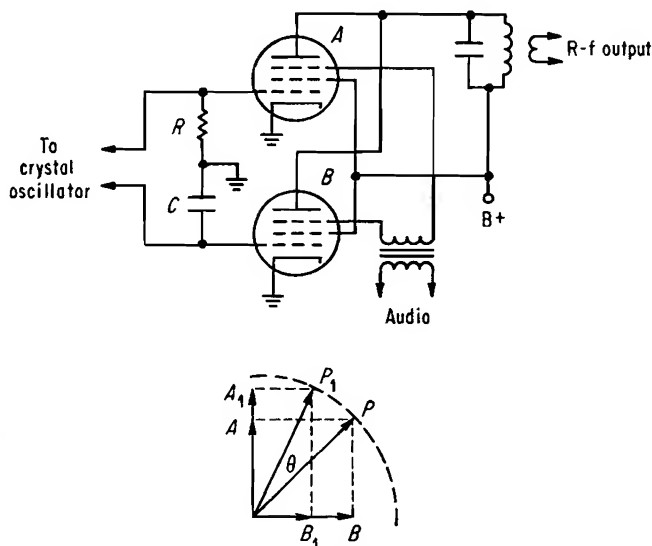


FIG. 7-4. Phase-modulation circuit.

must be used if distortion is to be avoided. The distortion will be less if the r-f voltages applied to the two grids are not equal because then the length of OP will remain more nearly constant during modulation. The distortion will be less than 2 per cent if the angle of swing is kept below $\pm 14^\circ$ (modulation index 0.25).

The system introduced by Armstrong (13,14), which was used in early f-m transmitters, is shown in the simplified block diagram of Fig. 7-5. The output of the crystal oscillator is connected to a 90° phase-shifting network, and also to a balanced modulator. The carrier is suppressed in the balanced modulator, and this has no output unless an audio signal is applied to the balanced modulator. The output of the phase shifter and the balanced modulator are then combined and multiplied. The result will be a p-m signal; in order to obtain frequency modulation, the audio input to the balanced modulator is passed through

a deemphasis network. In order to obtain 75-kc deviation at the transmitter output with a value of distortion less than 2 per cent at 30-cps modulation, it is necessary to multiply the modulator output frequency at least 5,000 times. For small values of modulation index, when there is only one pair of significant sidebands, the only difference between an a-m and a p-m wave is a 90° phase shift of the carrier with respect to the

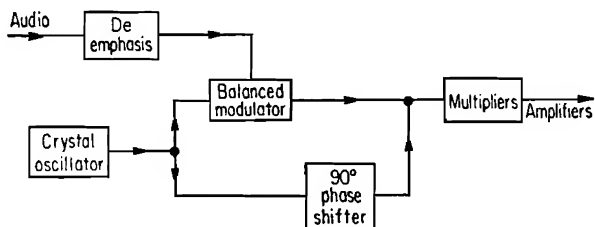


FIG. 7-5. Simplified Armstrong modulator.

sidebands. The Armstrong circuit can only be used when the modulation index is kept below 0.5 at the point of modulation. The modulation index can be increased to any desired amount by multiplication of the modulator output frequency.

A simple type of electronic phase shifter which is frequently used in mobile equipment is shown in Fig. 7-6. Variation of the transconductance of the tube causes variation in the phase angle between the input and output (15).

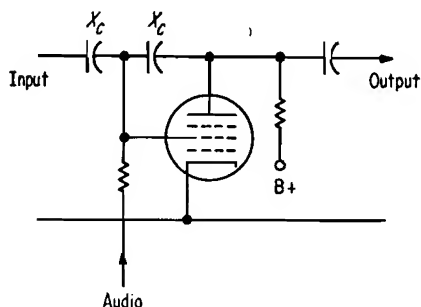


FIG. 7-6. Phase-shifting circuit.

ous value of the modulating signal. A block diagram is shown in Fig. 7-7. The output signal from the linear saw-tooth generator is limited at about 0.5 amplitude, and a pulse is derived at the start of the limiting (16). The limiting action is then bias-modulated by the modulating signal. The resulting pulses will be phase-modulated because their time of occurrence will vary.

Another p-m device which makes use of pulse techniques has been described by Gordon (17). In this circuit a multivibrator is controlled

¹ Radio Engineering Laboratories trade name.

by a crystal oscillator and the switchover period of the multivibrator is controlled by the modulating voltage. Less than 1 per cent distortion is obtained for 125° phase modulation at 400 cps.

A system of modulation described by Adler (18,19) uses a special vacuum tube which has a special deflector-grid assembly. A three-phase r-f voltage whose frequency is controlled by a crystal oscillator is applied to the 36-wire grid. Inside the tube a rotating ruffled electron disk is established. Outside the tube is located a modulating coil, which generates a magnetic field that causes the disk to be accelerated or

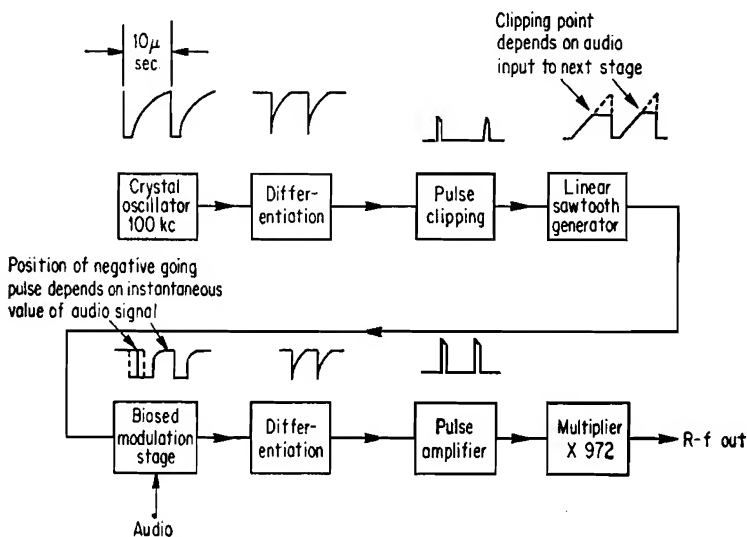


FIG. 7-7. Block diagram of Serrasoid modulator. (Courtesy of Electronics.)

retarded in accordance with the modulating voltage. The signal is picked up on an anode, and the output frequency is multiplied 432 times. After multiplication, an f-m signal with ± 75 kc deviation and less than 1.5 per cent distortion from 50 to 15,000 cps is obtained. A phase swing of ± 3.5 radians without exceeding 1.5 per cent distortion is possible at the tube. This tube is called a *Phasitron*.

7-6. Direct Generation of Frequency Modulation. When a modulator is used that produces the same amount of frequency deviation for different audio frequencies, the method is referred to as the *direct* method. In Sec. 7-5, phase modulators were discussed, and, when these are used in conjunction with an audio-correcting network to generate f-m signals, the method is referred to as the *indirect* method.

Mechanical Methods. One direct method of generating f-m signals is to connect a capacitor-type microphone or phonograph pickup directly across the tuned circuit of an oscillator. Similarly, the inductance of

the tuned circuit can be changed by varying the position of a powdered-iron slug which is close to the inductance. These methods are of value in *telemetry* applications where size and weight are of the utmost importance.

Reactance-tube Modulator. In this method the tank circuit of a Colpitts or Hartley oscillator is shunted by the plate-cathode circuit of a *reactance tube*, which is so arranged as to draw a reactive current that is varied in accordance with the modulating voltage. This reactive current has an effect equivalent to shunting a reactance across the oscillator-tuned circuit. The circuit is often referred to as a *quadrature* circuit because the r-f voltage developed across its output terminals leads or lags the r-f current. A simplified reactance modulator is shown in

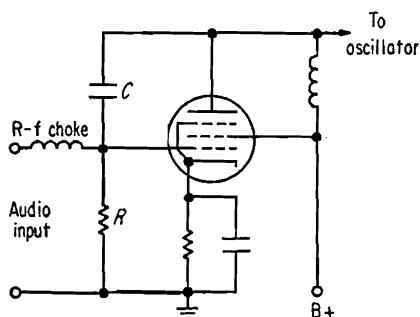


FIG. 7-8. Reactance-tube modulator.

Fig. 7-8. R and C constitute a phase-shifting network, and the voltage between the grid and ground can be made to be almost 90° out of phase with the voltage being supplied by the oscillator. The a-c component of the plate current will be opposite in phase to the grid voltage but will also be 90° out of phase with the oscillator voltage. The amplitude of the a-c component of the plate current will depend on the transconductance of the tube.

If the transconductance of the tube is varied by applying an audio voltage to the grid, the oscillator will be frequency-modulated. The circuit shown will inject capacitance into the oscillator circuit. If the R and C are interchanged, inductance will be injected into the circuit. Substituting an inductance for the C will have the opposite effect in each case. For the phase shift to approach 90° , the resistance should be less than one-fifth of the reactance when the resistance is between the grid and ground, and the reactance should be less than one-fifth of the resistance when the reactance is between grid and ground. For the circuit shown in Fig. 7-8, the frequency is approximately

$$f \approx \frac{1}{2\pi \sqrt{(C_o + g_m RC)L_o}} \quad (7-4)$$

where f = frequency, cps

C_o = oscillator capacitance, farads

R = resistance, ohms

C = capacitance, farads

g_m = transconductance, mhos

L_o = oscillator inductance, henrys

Input Capacitance Modulator. The grid-to-cathode capacitance of a vacuum tube depends on the plate-to-grid capacity, the transconductance, and the plate load. This is called the *Miller effect*. A simplified circuit is shown in Fig. 7-9. The grid-to-cathode capacity is given approximately by

$$C_{gk} = (1 + g_m R_L) C_{gp} \quad (7-5)$$

If a tube with a grid-voltage-vs.-transconductance curve having good linearity (such as the 6AK5) is used, a linear relationship between C_{gk} and the applied modulating voltage will be obtained. The plate circuit is tuned to the same frequency as the oscillator. The linearity is dependent on the plate tuning, and it may be found that the best linearity is obtained when the plate circuit is detuned slightly off resonance.

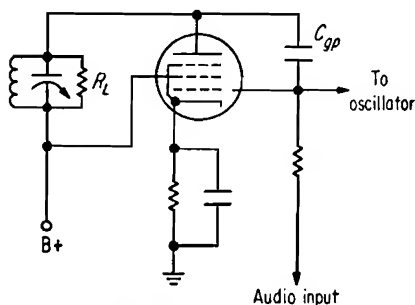


FIG. 7-9. Miller-effect modulator.

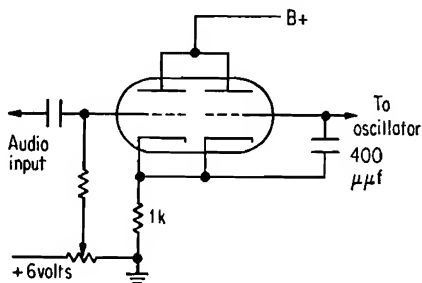


FIG. 7-10. Cathode-follower variable reactance. (Courtesy of Electronics.)

Cathode-follower Variable Reactance. A circuit which requires no tuned circuit, high reactances, or high resistances but which exhibits a relatively high Q is shown in Fig. 7-10 (20-22). The capacity injected by this circuit depends on the feedback reactance, gain, and output impedance of the amplifier. The triode with the grid-to-cathode capacity has its gain controlled by the other triode. With zero modulating voltage applied to the control triode, the injected capacity is about $250 \mu\text{f}$ for a 6J6 tube operated at 150 volts. A variation of ± 10 per cent for a grid-bias variation of ± 0.5 volt can be obtained with good linearity. Values of Q between 30 and 50 have been measured at 75 kc.

Modulation of Resistance-Capacitance Oscillators. If one of the resistance elements of an RC oscillator is replaced with a vacuum tube, the output frequency of the oscillator can be varied by changing the grid bias of the tube. By proper choice of constants and operating conditions the amplitude modulation can be reduced to a minimum, and harmonic distortion can be kept low even for swings up to ± 40 per cent of the

carrier frequency (23,24). This method is useful for modulation of subcarriers in facsimile or communication systems.

A simplified circuit of a modulated RC oscillator is shown in Fig. 7-11.

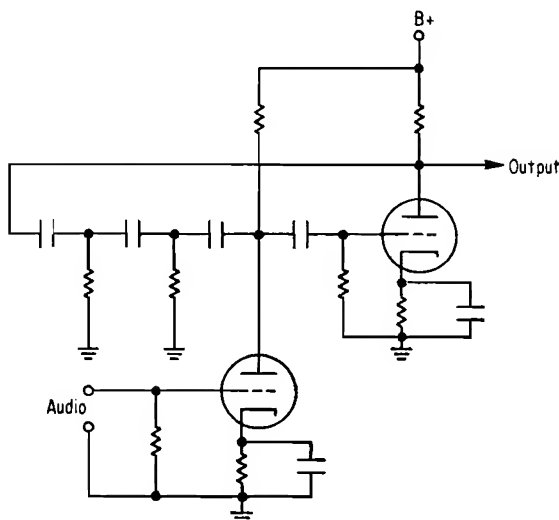


FIG. 7-11. Frequency modulation of RC oscillator.

Other Direct F-M Methods. A number of other methods of generating an f-m signal have been devised. One simple method is shown in Fig. 7-12 (25). Here the resistance of the diode is controlled by the modulator

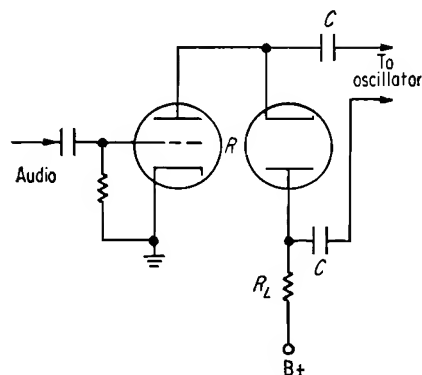


FIG. 7-12. Diode frequency modulator.

tube. As the resistance of the diode changes, the reactive current flowing through the capacitor C is changed, and consequently the oscillator frequency is changed. In another method the phase shift in the feedback circuit in an oscillator is changed at an audio rate. The phase-shift variation is independent of the modulating frequency, and therefore direct frequency modulation is produced (26). The frequency deviation of

a modulated oscillator can be increased by inserting a stage of amplification between the oscillator-tuned circuit and the grid of the reactance tube (27). It has also been shown (28) that the deviation can be increased by inserting a cathode-follower stage between the reactance modulator and the oscillator. The use of

phase-shifting circuits can be avoided by inductively coupling the modulator to the oscillator (29). Inductive coupling has also been used to increase the modulating effect of a condenser microphone which is directly modulating an oscillator (30). The oscillator and modulator functions can be combined in one tube (31). In another circuit the modulator and midfrequency stabilization are combined (32). Frequency modulators have been analyzed in detail in a number of places in the literature (33-39).

Frequency Modulation of Klystrons.

A reflex klystron can be frequency-modulated directly by applying the modulating voltage to the repeller electrode. (See Sec. 4-7 for a discussion of this type of klystron.)

7-7. Pulse Modulation. Pulsed transmissions are used in radar, navigation, and communication systems. In radar systems the pulse characteristics generally do not vary, and information is derived from the system by measuring the time interval between transmission and reception. In navigation and communication systems, some characteristic of the pulse transmission is varied or controlled so that information can be extracted from the pulses. The characteristics of a pulse that can be controlled are the amplitude modulation (PAM), duration or width modulation (PDM or PWM), and the position, time, or frequency modulation (PPM, PTM, or PFM). The three basic systems are illustrated in Fig. 7-13.

In another method called pulse-count or pulse-code modulation (PCM), coded groups are sent which correspond to the amplitude of the signal being transmitted at the instant a sample is taken. The signal which is being transmitted by pulse-count modulation is broken down or *quantized* into a number of discrete steps of amplitude for conversion into coded signals. The coded pulses are usually transmitted in a *binary* code, which means that the pulse has only *two* possible states. These states will be on and off in the case of an a-m signal or one of two fre-

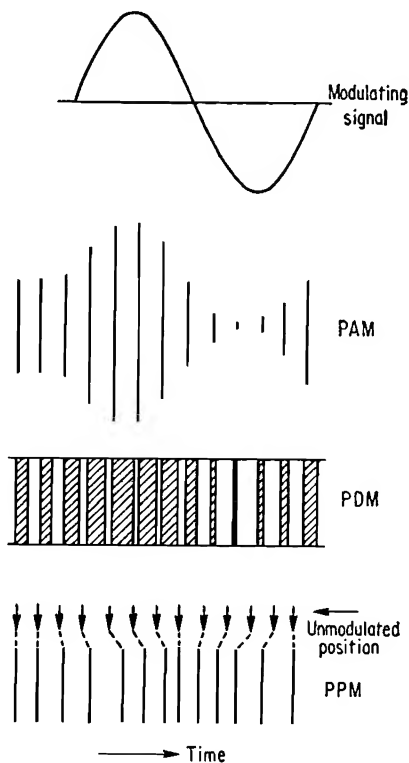


FIG. 7-13. Three types of pulse modulation.

The three basic systems are illus-

quencies in the case of an f-m signal. It is also possible to use pulses with *three* stable states; the code in this case would be called a *ternary* code. The binary code may have any number of digits or *bits* per group. The number of bits in the code will determine the number of discrete values used for quantizing the signal. The relation is

$$\text{Number of quantum levels} = 2^n \quad (7-6)$$

where n is the number of bits in the code. For example, if there are 3 pulses per group, the signal being transmitted will be quantized into

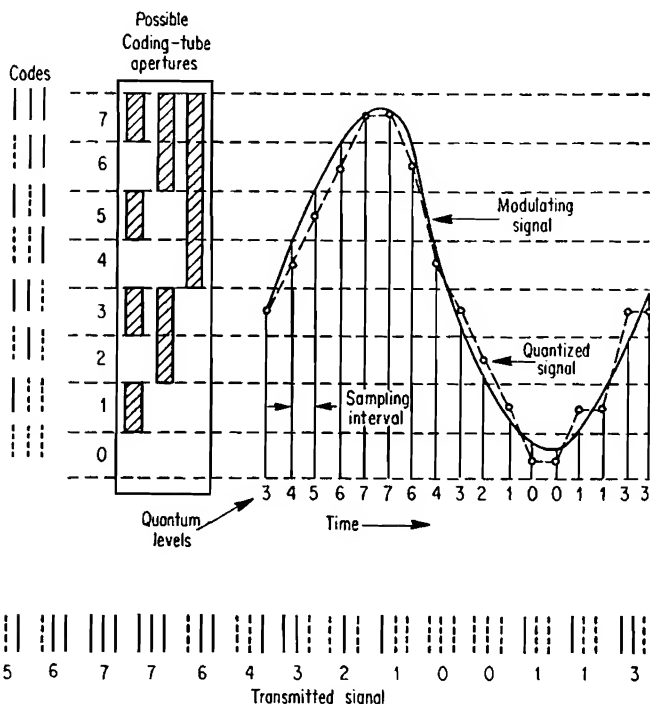


FIG. 7-14. Pulse-code modulation. (From H. S. Black, "Modulation Theory," D. Van Nostrand Company, Princeton, N.J., 1953.)

2^3 or eight levels. An example of 3-bit coding is shown in Fig. 7-14. The advantage of pulse-code modulation is that the signal-to-noise ratio in the received signal is not degraded when it is passed through a large number of amplifiers or repeaters (40,41).

Millman and Taub have given an excellent treatment of pulse circuits (42).

7-8. Sidebands Produced by Pulse Modulation. A pulsed transmission such as is used in radar or navigation transmitters, also has a

sideband distribution depending on the transmitter characteristics. In the case of radar sets, the *duty cycle* or ratio of pulse-on time to pulse-off time may be as small as 0.001. An examination of such a signal on an oscilloscope with a linear time base would yield little information other than the time of occurrence. On the other hand, a spectrum analyzer will display an important indication of the performance of the output stage. Figure 7-15 shows the output spectrum of a platinotron modulated with 17- μ sec pulses. Since the duty cycle is relatively small, there are not enough pulses per sweep of the oscilloscope to fill up the display, and there are actually more sidebands than vertical lines displayed by the spectrum analyzer. The usual objective in adjusting a pulsed transmitter is to send the desired information while the spectrum occupies a minimum amount of bandwidth (43-45).

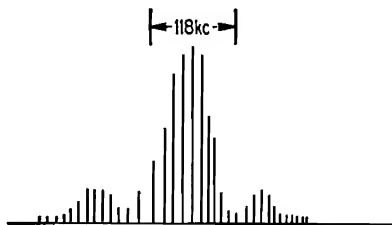


FIG. 7-15. Sideband spectrum of pulse-modulated signal. (Courtesy of Electronics.)

7-9. Pulse Generators. There are two basic methods of generating pulsed transmissions. The first is to generate pulses at low level and to amplify. The second method is to control a high-voltage power supply so as to switch on a high-powered oscillator such as a magnetron for brief intervals.

Low-level Pulse Generation. There are two principal methods of generating pulses at low levels:

1. *Differentiation of square waves*
2. *Blocking oscillators*

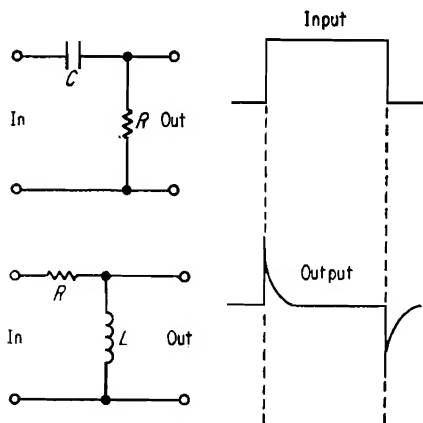


FIG. 7-16. Differentiating circuits.

Simple *RC* and *RL* differentiating circuits are shown in Fig. 7-16. The leading edge of the output pulses will be the same as the leading edge of the square wave used for the input. The trailing edge will depend on the *RC* or *RL* time constant. The time required for the trailing edge of the pulse to fall to 37 per cent of its initial value or to reach 63 per cent of its final value is *RC* or *L/R* sec where *R* is in ohms, *C* in farads, and *L* in henrys. Figure 7-17 gives a universal time-constant chart for *RC* and *RL* circuits.

Square waves may be generated by two principal methods:

1. *Multivibrator* circuits
2. *Clipping* a sine wave

There are many kinds of multivibrator circuits. It is easy to obtain less than 1 μsec rise time and 0.05- μsec rise times are possible.

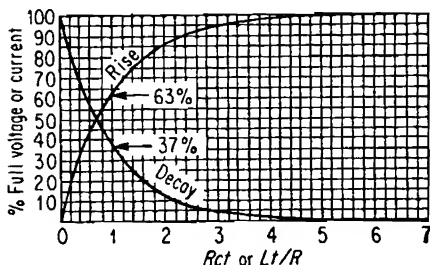


FIG. 7-17. Time constants for RC and RL circuits.

Three general types are the *monostable*, *bistable*, and *astable* multivibrators (46-48). The monostable or one-shot circuits have one stable state and one unstable state. When a trigger is applied, the circuit will change to its unstable state and after a time interval determined by its time constants will return to its stable state. A rectangular pulse of a definite length will thus be generated each time a pulse is applied. This is also called a *delay* multivibrator. The bistable multivibrators have two stable states and will change from one to the other each time a trigger is

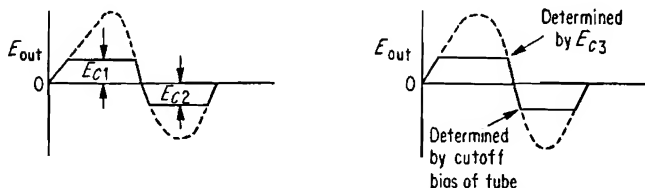
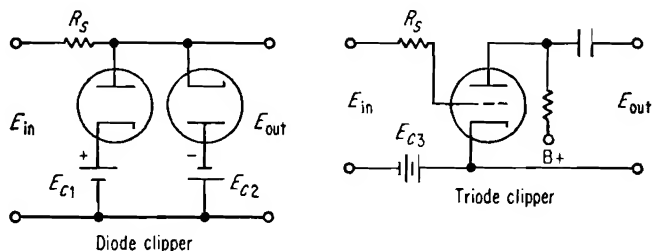


FIG. 7-18. Diode and triode clippers and their waveforms.

applied. These are also called *scaling* or *scale-of-two* circuits. The bistable multivibrators are also called *flip-flop* circuits, although sometimes the *monostable* circuits are referred to as flip-flops. Astable multivibrators have no stable state and are free-running or self-oscillating.

A type of circuit which is similar to the multivibrator but which uses a linear timing waveform with increased timing stability, is called a *phantastron* (49).

In clipping, a portion of the sine wave is *limited* to some arbitrary level, which will be independent of the level of the original signal. The positive or negative halves of the wave, or both, may be clipped. Clipping may be accomplished with biased diodes or by *grid limiting* and *cutoff limiting* in a tube (50). Figure 7-18 shows simple diode and triode clippers.

A blocking oscillator is a tuned-plate oscillator with circuit constants proportioned so as to cause an extreme case of intermittent oscillation (51). A typical circuit is shown in Fig. 7-19. L_p and R_g are made large while C_g is made small. The pulse length and shape is largely determined by the characteristics of the transformer.

All of the above types of low-level pulse generators can be synchronized with a stable oscillator.

High-peak Power Pulses. There are two basic types of power-pulse generator (52,53):

1. *Hard-tube* pulser or modulator
2. *Line-type* pulser

When the hard-tube pulser is used, only a fraction of the stored energy in the power supply is discharged into the load. With the line-type pulser all the stored energy in the power supply is discharged during each pulse.

To accomplish the energy discharge into the load, it is necessary to provide a suitable switch that can be closed for the length of time corresponding to the pulse duration and maintained open during the time required to build up the stored energy again before the next pulse. In the hard-tube pulser the energy is stored in a capacitor, and the switch is a high-vacuum tube with a control grid. A gaseous-discharge device such as a thyatron cannot be used in this circuit because, once a thyatron is fired, the conduction cannot be stopped with the control grid but can be stopped only by removing the plate voltage.

In the line-type pulser the energy is stored in a lumped-constant transmission line. In the usual type the energy is stored in the capacitances, and the networks are called *voltage-fed* types. In this type of pulser the switch used for discharging the line into the load may be a thyatron because the energy in the line is completely discharged. This permits the plate voltage on the thyatron to drop to a value low enough for

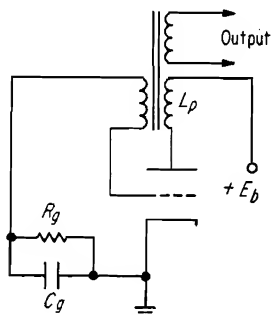


FIG. 7-19. Blocking oscillator.

conduction to stop. Since the shape and amplitude of the pulses will depend on the circuit values in the line, it is usually called a pulse-forming network (PFN). A pulse-forming network is also frequently used in the driver stage for hard-tube pulsers. Spark gaps may be used for the discharge of line pulsers.

A simplified circuit for a hard-tube pulser is shown in Fig. 7-20 and one for the line-type pulser is given in Fig. 7-21.

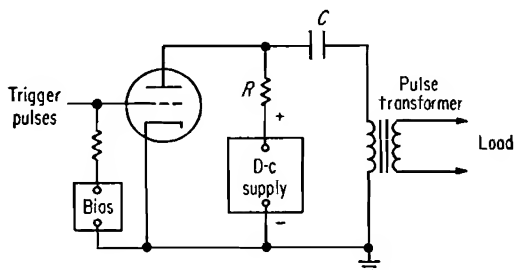


FIG. 7-20. Hard-tube pulser.

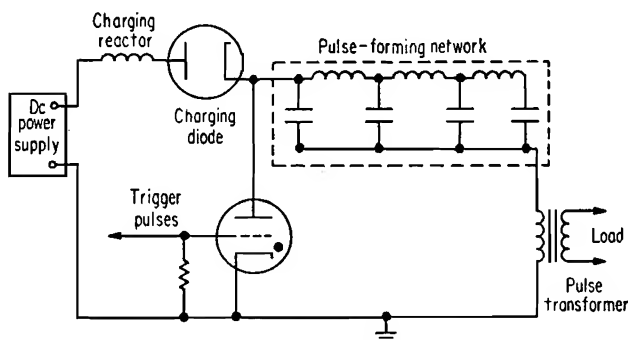


FIG. 7-21. Line-type pulser.

Grid Pulsing. The line-type and hard-tube pulse modulators are used for pulsing tubes without grids such as magnetrons, platinotrons, and traveling-wave tubes. It is also convenient in some cases to plate-pulse triodes or tetrodes. If, however, the application allows the use of a triode or tetrode, or a klystron with a control electrode, the need for a pulse modulator may be avoided by biasing the grid to cutoff and then pulsing the tube up out of the cutoff condition.

7-10. Pulse-amplitude Modulation. A PAM r-f signal can be generated by pulsing the r-f drive to a plate- or grid-modulated amplifier. Also the r-f drive may be constant, and the plate or grid may be modulated with pulses whose amplitude varies with the modulating signal. It is more usual to modulate the pulses because most modulated-pulse transmissions are used in systems where several *multiplexed time-division*

channels are used. Multiplexing will be discussed in detail in Chap. 12. The a-m pulses may, in turn, either amplitude- or frequency-modulate the r-f carrier. Figure 7-22 shows a simplified generator of PAM signals.

7-11. Pulse-duration Modulation. This type of modulation is also called pulse-width or pulse-length modulation. The time of occurrence of the leading edge, trailing edge, or both edges of the pulse may be varied. Pulse-duration is also considered to be included within the

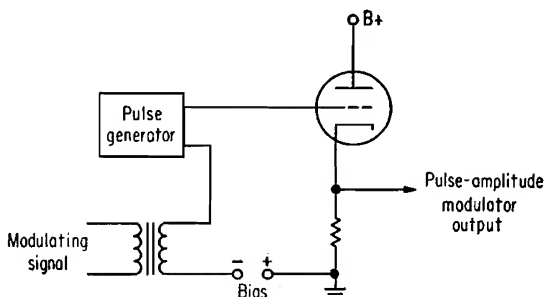


FIG. 7-22. Pulse-amplitude modulator.

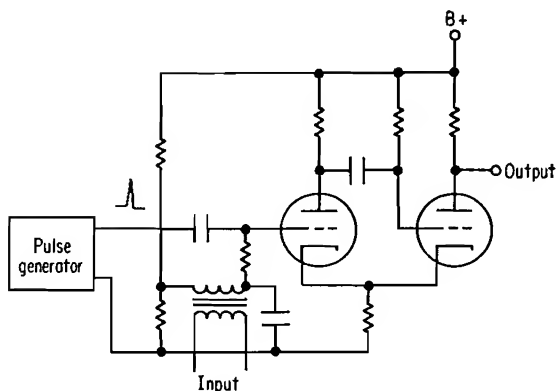


FIG. 7-23. Pulse-width modulator.

general classification of pulse-time modulation. Pulse-time modulation includes those types in which any characteristic of the pulse has its time of occurrence varied.

Pulse-duration modulation (PDM) has an advantage analagous to the advantage of frequency modulation over amplitude modulation. Noise and interference can be reduced at the expense of increasing the bandwidth occupancy, provided that the noise level is below a certain threshold.

There are several methods for producing pulse-width modulated signals (54,55). One possibility is shown in Fig. 7-23. In this circuit, trigger pulses and the modulating signal are both applied to a monostable or

one-shot multivibrator. Each trigger pulse initiates a cycle of multivibrator operation, and the multivibrator switches back to its stable state after a time interval that varies linearly with the modulating voltage. The leading edges of the output pulses occur at fixed time intervals while the time of occurrence of the trailing edge will vary.

7-12. Pulse-position Modulation. In pulse-duration modulation no information is actually being transmitted between the beginning and end of each pulse, and power will obviously be saved if the transmitter is turned off during this interval (56). Also, only the edge which is being changed needs to be transmitted. If a PDM signal is merely passed through a differentiating circuit, PPM signals will be obtained. A diode can then be inserted in the circuit to discard either the leading or trailing edge of the original PDM signal.

Pulse-position modulation is another special example of pulse-time modulation. If the pulse-repetition rate (PRF) of the pulses is varied, the signal will be somewhat similar but the position of the pulses can not be referred to a fixed time or position reference. This is called pulse-frequency modulation (PFM) (57). This is not to be confused with pulsed frequency modulation which refers to an f-m carrier that is pulsed with constant-amplitude-constant-duration pulses (58).

7-13. Pulse-code Modulation. Pulse-code modulation has also been called *pulse-count* modulation. In the previously discussed methods of modulation, the quantity modulated is continuously variable over the entire range. In pulse-code modulation the information to be sent is subjected to two successive transformations. First the information is periodically sampled and then quantized, which means that the amplitude of the sample is approximated to the nearest of a finite number of discrete levels. Then the level that is obtained is converted into a number of pulses and spaces according to a predetermined code.

When the pulse signals are above the noise by a specific threshold value, only the presence or absence of each signal need be determined at the receiver, and the noise will be considerably reduced (59). On the other hand, some distortion or noise will be introduced because of the difference between the quantized and exact amplitude samples. This distortion will decrease as the number of quanta or discrete levels is increased (60,61). This is shown in Fig. 1-1.

Translation of the quantized signal into the code usually requires a special type of tube. One type is similar to a cathode-ray tube with the fluorescent screen replaced by an aperture plate. The pattern of holes in the aperture plate corresponds to the code for that amplitude (62,63). Another system uses a tube with ten separate targets and a second tube in which the electron beam has 10 stable positions produced by crossed electric and magnetic fields (64).

These special tubes perform both the quantizing and coding. PAM signals are applied to the vertical deflection plates. A sweep signal is applied to the horizontal-deflection plates, which sweeps the beam across the tube once for each sample that is applied to the vertical plates. As the beam sweeps across the aperture plate, the beam current that passes through the holes is collected sequentially by the collector plates. The number of plates will correspond to the number of bits in the code. If

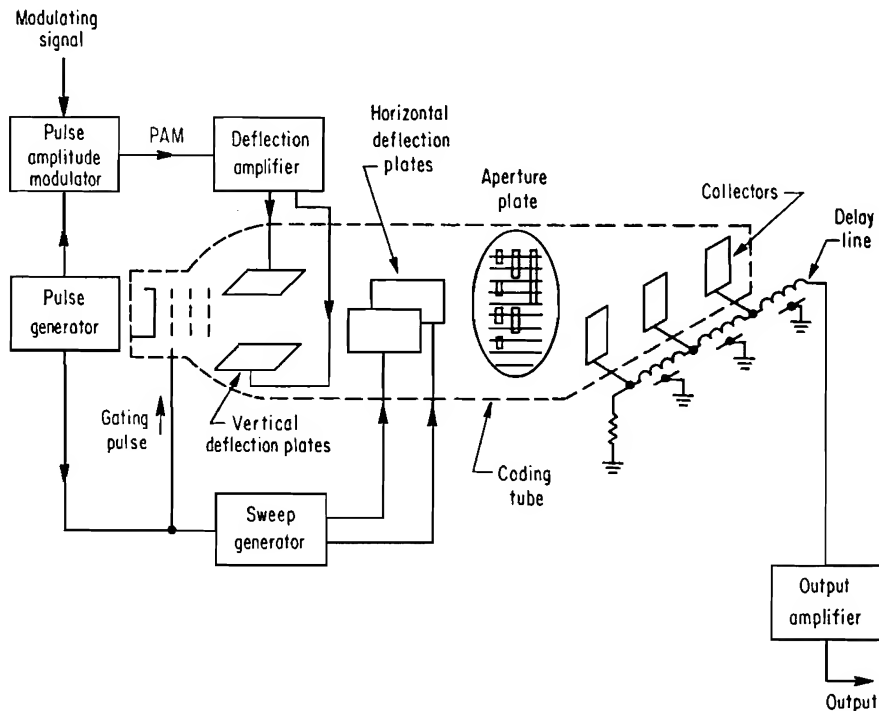


FIG. 7-24. Pulse-code modulator. (Courtesy of IRE.)

there are 5 bits, there will be five plates, and the number of vertical levels will be 2^5 or 32. The five plates may be connected to taps on a delay line so that the pulses may be spread out in time. (Figure 7-24 shows a block diagram of a PCM modulator.)

7-14. Delta Modulation. A type of PCM which does not require a special coding tube is called *delta* modulation (65,66). This system employs a train of equally spaced identical pulses, which are transmitted only when the magnitude of the output from a local receiver, within the transmitter, falls below the modulation signal to be transmitted.

In other words, pulses are transmitted when the modulating signal is

rising but are not transmitted when it falls faster than a certain rate determined by time constants of the circuit.

The local receiver is placed in a feedback loop in the transmitter. It is supplied with the transmitted pulses and pulses of the opposite polarity, whenever the transmitted pulses are absent. The received pulses generate

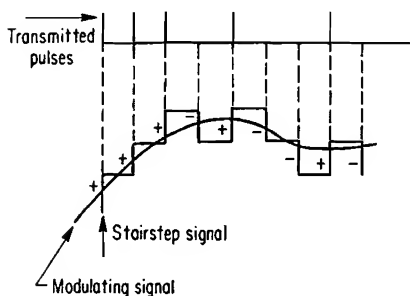


FIG. 7-25. Delta modulation staircase signal. (Courtesy of Electronics.)

a staircase signal which approximates the modulating signal as shown in Fig. 7-25. The staircase signal is compared to the modulation signal in a difference circuit. When it is greater than the modulation, no pulses are transmitted, and pulses of opposite polarity reach the receiver. When it is less than the modulation, pulses are transmitted and reach both the local and any remote receiver. The train of

transmitted pulses will be a measure of the modulating signal. Satisfactory voice transmission is possible when pulses are transmitted at a rate of 60 kc or more.

A block diagram of the modulator is shown in Fig. 7-26. When gate 1 is open, positive pulses are transmitted and also reach the local receiver. The positive staircase output from the receiver is compared with the

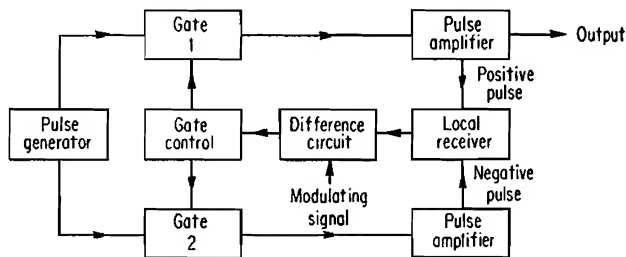


FIG. 7-26. Delta modulator. (Courtesy of Electronics.)

modulating signal in the difference circuit, and, if the difference is negative, gate 1 remains open. If the difference becomes positive, gate 1 is closed, and gate 2 is opened, allowing negative pulses to reach the receiver which will cause the difference to approach zero.

The frequency response is limited by the fact that the number of levels available for approximating a given signal is limited by the frequency and amplitude of the pulses. The number of levels or quanta available at the receiver is approximately given by

$$\text{Number of levels} \approx \frac{f_c}{2\pi f_m} \quad (7-6)$$

where f_c = pulse-repetition rate, cps

f_m = modulating frequency

For example, if the pulse-repetition frequency is 100 kc and the modulating frequency is 1,000 cps, the number of levels available at the receiver (ratio between pulse amplitude and peak audio amplitude) will be about 16. As the modulating frequency decreases, the amplitude that may be transmitted increases rapidly since the number of available levels increases.

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POWER SUPPLIES

8-1. A-C Voltage Regulators. General. A-c voltage regulators are used to stabilize the transmitter operation against incoming line-voltage variations. In so doing, the a-c regulator serves to prolong the life of the tubes and components by maintaining the proper operating conditions within the transmitter. Regulation of the filament voltage in a transmitter will help to ensure maximum tube life. The effect of the filament voltage on tube life was discussed in Chap. 4. Any increase in the line voltage will also result in increased plate dissipation in the transmitter tubes. The plate dissipation will increase as the square of any line voltage increases. While the increased plate dissipation may not result in immediate damage, it may serve to shorten tube life. For example, a 15 per cent increase in the line voltage will increase the tube input power by 32.25 per cent. The effect of the line-voltage increase goes beyond its effect on the transmitter tubes. Since this increased power consumption is supplied by the power supply, all these components must be capable of supplying the increased load. This includes the rectifiers, filter capacitors, chokes, and transformers. If the components are marginally rated, shortened component life will be the inevitable result. This is particularly true in transformers where the transformer life is largely dependent on the insulation operating temperature.

There are three general types of a-c line-voltage regulators in general use. Each of these types has several variations, depending on the manufacturer of the item. Specifically, these types are, (1) the electromechanical regulator, (2) the resonant transformer regulator, (3) the saturable-reactor regulator.

The Electromechanical Regulator. As the name implies, the electromechanical regulator achieves its regulation through a mechanical change in position of a voltage-compensating device. This is usually driven by an electric motor energized by a servo system that samples the regulator output voltage. If a change takes place in the regulator a-c output voltage, an error signal is sent to the servomotor which rotates the voltage-compensating device to correct for the voltage change.

The type shown in Fig. 8-1 uses a brush type of variable autotransformer driven by a servomotor to accomplish the voltage change. The variable autotransformer adjusts the primary voltage to an auxiliary voltage-boosting transformer. The secondary of this transformer is in series with the line and load. Thus as the primary voltage is varied by the variable autotransformer, the secondary voltage, and hence the regulator output voltage will vary accordingly.

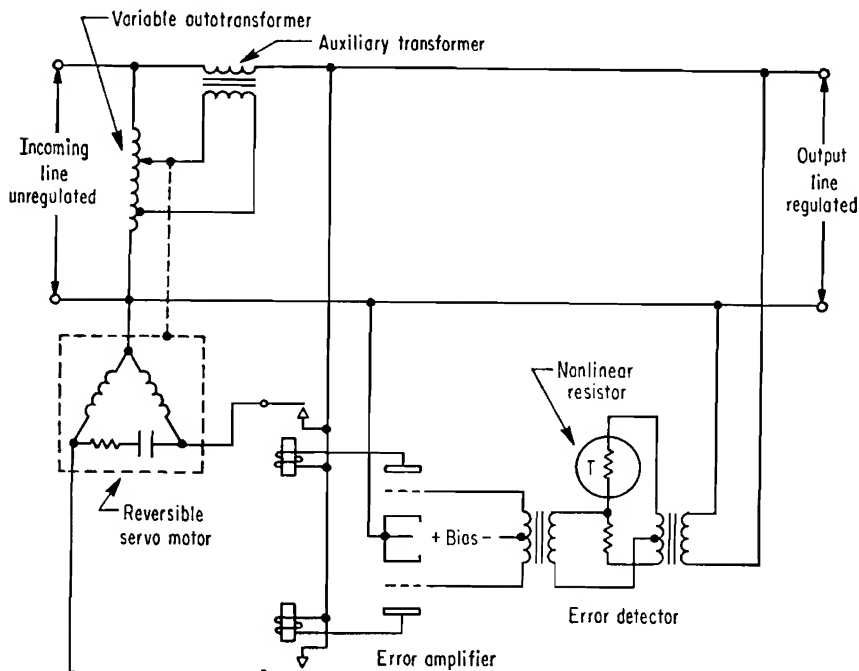


FIG. 8-1. Simplified circuit for an electromechanical regulator using a variable autotransformer.

The error detector, which continuously samples the regulator output voltage, makes use of a current- and temperature-sensitive resistor in a bridge circuit. Any change in the regulator output voltage will change the current through the nonlinear resistor. This changes the resistance of the nonlinear resistor and hence unbalances the bridge. Thus an a-c error signal is sent to the servo amplifier which closes a relay which will rotate the motor in the proper direction to compensate for the change in the output voltage.

This type of regulator has the slight disadvantage of having a moving part which requires some occasional maintenance. This applies particularly to the brush of the variable autotransformer. This type

usually makes use of tubes which inevitably require some maintenance. However, compared to the electromechanical induction regulator to be described, it provides a lower cost per delivered kva.

A second type of electromechanical a-c line-voltage regulator, commonly known as an *induction regulator*, is shown in Fig. 8-2. This type of regulator uses a rotor and stator construction similar to that found in

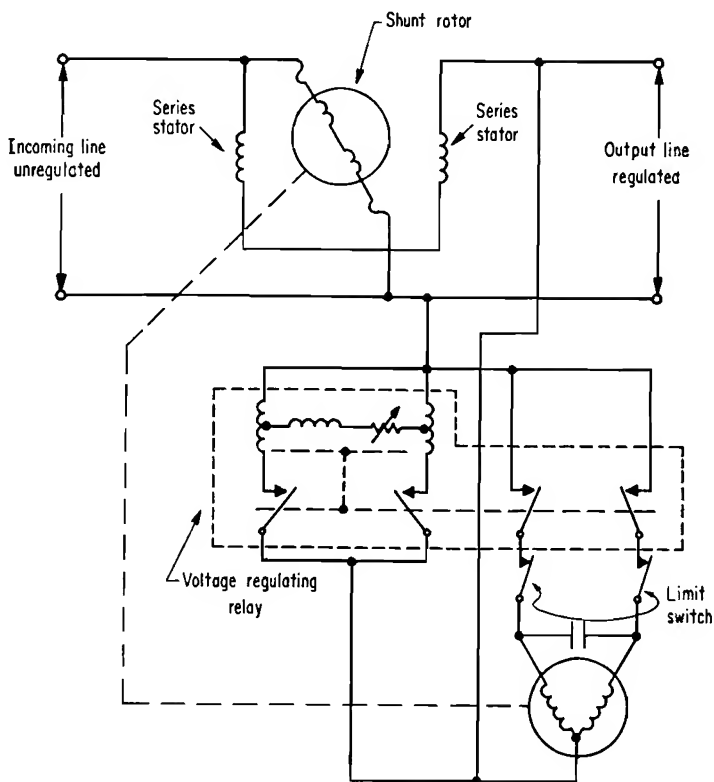


FIG. 8-2. Simplified circuit of an electromechanical regulator using a variable induction transformer.

motors, except that in this case the rotor is capable of rotating through only a fraction of a revolution. Its position is changed by a servomotor driven through a reduction gear. The stator winding is energized by the incoming line. The rotor is in series with the incoming line and the load. A voltage is induced into the stator from the shunt-connected rotor, which adds vectorially to the line voltage. The magnitude of this voltage is determined by the physical position of the rotor with respect to the stator.

The voltage-sensing system of the induction regulator does not use vacuum tubes. Instead it uses a special voltage-sensing relay which is placed directly across the regulator output. Any change in the regulator output voltage due to a change in the incoming line voltage or a change in the load will cause the motor to rotate the induction regulator to compensate for this voltage change. The sensitivity of the relay to voltage changes can be adjusted. While this type of regulator is rugged and dependable, it has the disadvantage of being rather costly per regulated kva output.

The Resonant Transformer Regulator. This type of regulator is shown in its basic form in Fig. 8-3. The variable incoming line voltage is connected to the regulator transformer primary. This produces a magnetic flux ϕ which cuts the compensating and secondary windings. The capacitor across the compensating winding resonates with this winding and the leakage reactance of the secondary winding. A partial magnetic shunt is arranged between the primary winding and the compensating and secondary windings. When resonance takes place, the iron core for the compensating and secondary windings becomes saturated. The magnetic shunt prevents the magnetic circuit associated with the primary from saturating. The shunt, core, and winding are so proportioned that any flux changes in the primary magnetic circuit

resulting from changes in the primary voltages will have no effect on the flux in the secondary magnetic circuit due to the resonant condition. This latter section remains fully saturated. Thus the resultant constant value of flux in the secondary magnetic circuit induces a constant voltage in the secondary winding that supplies the output line.

This type of regulator has the advantage of having no moving parts. Its simple construction also enables this form of regulator to be economical, particularly in the smaller sizes where the core size is reasonable. It is also capable of providing a fairly rapid response to variations in the input voltage compared to the electromechanical regulators previously described. However, it suffers from the disadvantage that it does not completely regulate against changes in the load. Any change in the load current will result in a change in the regulating point. This is illustrated in Fig. 8-4. Any change in the power factor of the load will also affect

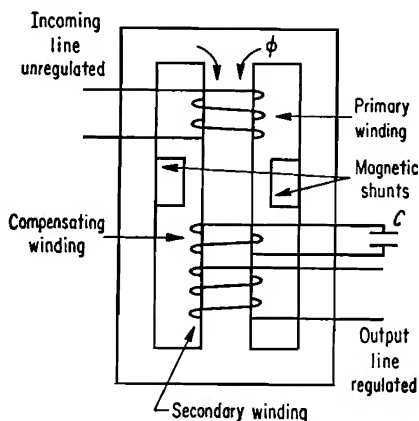


FIG. 8-3. Simplified circuit of a resonant transformer a-c regulator.

the regulating point of the regulator. This is illustrated in Fig. 8-5. However as long as the power factor remains fixed, then the regulator will hold against input-voltage variations. Since most regulators of this type are designed to work into resistive loads, some care must be used in

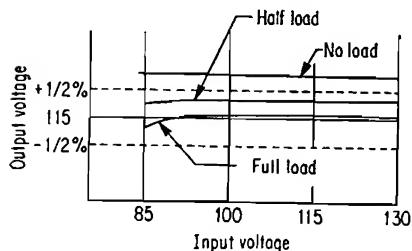


FIG. 8-4. Regulation curve of a typical resonant transformer regulator. Note the change in the output voltage with change vs. regulator load.

attempting to regulate against input-voltage variations with a capacitor input filter power supply as the load. However, the regulator can be specially designed to compensate for the poor power factor of the load, or, if desired, some form of power-factor correction can be used.

This type of regulator also suffers from severe harmonic distortion of the output waveshape. This distortion may be as much as 10 to 20 per

cent of the total output voltage unless some form of filter is added to minimize this distortion. This distortion is due to the saturation of the iron core, which is the basis of this method of regulating.

The Saturable-reactor Regulator. This type of regulator utilizes the controlled variable impedance of a saturable reactor to achieve the

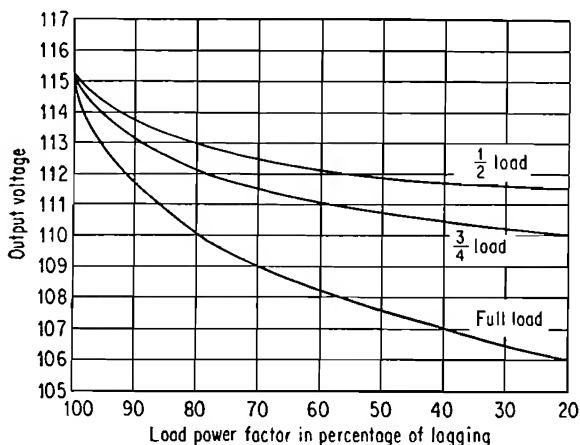


FIG. 8-5. The effect of the load power factor on the output voltage of a typical resonant transformer regulator.

voltage control. The saturable reactor is controlled by a d-c current flow developed by an a-c error-voltage detector and amplifier. Operation of this error detector will vary according to the manufacturer of the regulator; however, a typical method is shown in Fig. 8-6. In this circuit, a diode is used as one leg of a Wheatstone bridge. The emis-

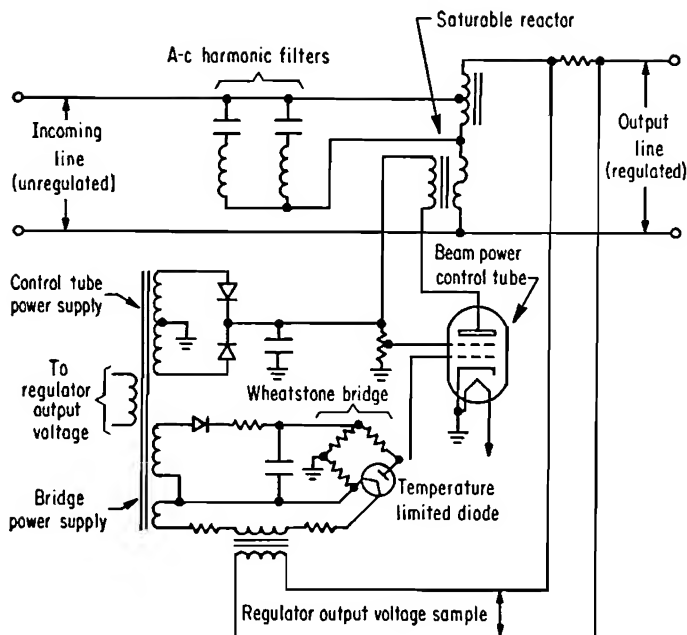


FIG. 8-6. Simplified circuit of a saturable reactor electronic regulator.

sion of the diode is limited by the diode filament temperature and specifically by the applied filament voltage. The filament voltage for this diode is obtained from the regulator output. Thus, if the regulator output changes, the bridge will unbalance owing to the resultant filament-voltage change of the diode. The change is then applied to the grid of the amplifier control tube. This results in a change in the direct current through the saturable-reactor control winding and so changes the impedance of the reactor. In the circuit shown, the reactor is in series with an autotransformer across the a-c line. Thus any impedance change in the saturable reactor will result in an output-voltage change. The current

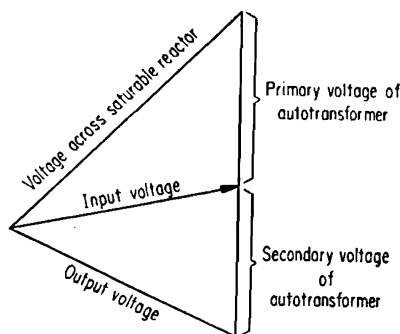


FIG. 8-7. Phase relationships in the saturable reactor regulator shown in Fig. 8-6.

change through the saturable reactor, and hence the voltage change in the autotransformer is such that the output voltage is restored to its former value, and the bridge is essentially brought back into a balanced condition. The phase relationships that exist in this type of regulator is shown in Fig. 8-7. Other voltage-error-detecting systems utilize

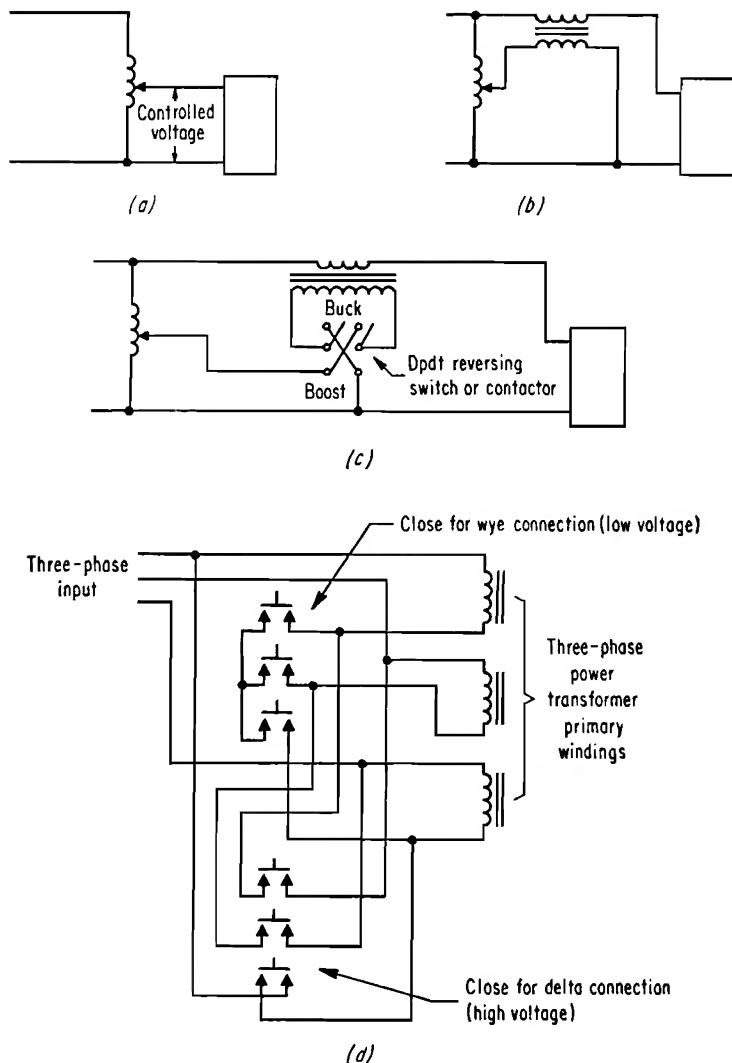


FIG. 8-8. Various methods of manual a-c voltage control. (a) With variable autotransformer—full control; (b) using variable autotransformer to supply auxiliary buck or boost transformer—partial control; (c) same as *b* except autotransformer primary can be reversed by switch to permit voltage buck or boost; (d) transformer primary switched from delta to wye for high-low voltage control; (e) by means of a variable-induction transformer; (f) by means of a series saturable reactor.

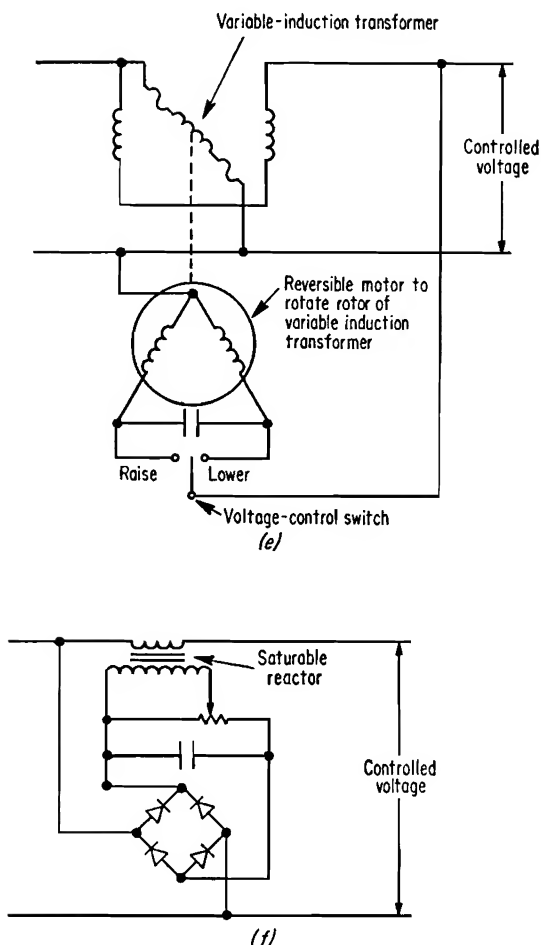


FIG. 8-8. (Continued)

Zener voltage-reference diodes and additional stages of magnetic amplifiers to achieve control of the saturable reactor. However, the basic principle of regulation is still the same.

The saturable-reactor regulator provides a rapid response to voltage changes and generally provides regulation to better than 0.1 per cent. However, this regulator is generally quite costly per regulated kva and tends to produce some amount of harmonic distortion because of the saturation of the reactor. Most manufacturers include some additional filtering to minimize the harmonic output. This type of regulator is generally available with outputs from 1 to 15 kva.

8-2. Power-supply Voltage Control. In many types of equipment it is very desirable to manually control the d-c voltage applied to the tubes.

This is most conveniently done by controlling the a-c line voltage applied to the primary of the d-c power supply. Many of the same voltage-controlling elements used in the automatic voltage regulators are readily adapted to the purpose of manually controlling the a-c input voltage. However, while the principle may often be the same, the size of the controlling device may differ considerably. Since the range of the manual voltage control is generally greater than most of the automatic voltage regulators, the size and electrical capacity will increase accordingly.

Some commonly used methods of voltage control are illustrated in Fig. 8-8. The variable autotransformer of Fig. 8-8a is capable of varying the power-supply voltage from zero up to the maximum voltage of the supply. This may not always be necessary nor desirable. Therefore some saving in the cost of the components is sometimes possible (particularly in larger transmitters) by limiting the voltage control range and by using the bucking and boosting arrangement shown in Fig. 8-8b. This enables the use of a smaller and lower-rated variable autotransformer. If the primary of the auxiliary transformer shown in Fig. 8-8b is reversed by means of a switch or relay, then the output line voltage can be bucked or boosted by the magnitude of the auxiliary transformer secondary voltage. This arrangement will give twice the range of voltage control, with the same kva-rated components. This is shown in Fig. 8-8c.

In situations that need only a reduced-voltage position for tuning and initial adjustment, the delta-wye transformer primary switching system shown in 8-7d can be used. This is applicable only to power supplies using a three-phase power transformer. With the switch or contactor in the wye primary position, the voltage will be 57 per cent less than when the primary is connected in delta, or the full-voltage position. This system requires that all the leads from the primary be brought out of the transformer and not connected internally.

This method is often combined with the buck-boost arrangement of Fig. 8-8c in a three-phase system to provide four voltage ranges.

The method of Fig. 8-8e utilizes a manually controlled induction regulator similar to that previously described. In this case the induction regulator can be rotated either by hand through a reduction worm gear or through a manually controlled reversible motor.

Figure 8-8f illustrates the use of a series saturable reactor for controlling the a-c voltage applied to a power supply. This latter requires the use of a stable source of d-c to control the reactor. The controlling power is usually 5 to 10 per cent of the power controlled by the reactor.

KVA Rating of Buck-Boost Transformers and Induction Regulators. Since the buck-boost transformer and the induction regulator are both

used in automatic voltage-regulating systems and in manually controlled voltage systems, it is worth noting the factors that control the ratings of these components. Both these components must have a secondary winding capable of carrying the full load current of the d-c supply. The secondary winding must also be capable of developing the required bucking or boosting voltage. This will depend on the voltage control range desired. For example, if a 50 per cent voltage control range is desired, then this transformer, if it were to buck the line voltage only, would have a secondary voltage equal to 50 per cent of the line voltage. If it were arranged to buck as well as boost, then the secondary voltage need only be 25 per cent of the nominal line voltage to achieve a total 50 per cent voltage change. The rating of the transformer or induction regulator would then be found from the equation

$$E_{\text{sec}} I_{\text{max}} = \text{kva rating} \quad (8-1)$$

where E_{sec} = maximum buck-boost transformer voltage

I_{max} = maximum a-c line current through secondary of buck-boost transformer

8-3. Power-supply Circuits. The basic power-supply configurations most commonly used in transmitter circuitry are illustrated in Table 8-1. This figure also gives all the pertinent operating and design considerations. The data given in Table 8-1 are based on a sine-wave input and an infinite inductance choke and assume no transformer or rectifier losses. The rectifier voltage drop will vary according to the rectifier type selected, and at the higher voltages it can be ignored since the rectifier drop is small compared to the total output voltage. Since these are all basic circuits, the filament connections (if there are any) for the rectifiers are not shown. In many cases, the rectifier filaments can be operated in parallel, thus reducing the number of transformers needed. For example, the three-phase full-wave circuit will require a minimum of four rectifier transformers for the six rectifiers. Three of the rectifiers have a common-cathode connection and therefore are at the same operating voltage, while three of the rectifiers must be isolated since the cathodes are at different operating voltages.

However, the use of separate transformers for each rectifier filament has the advantage that only one type of transformer need be kept as a spare, while the filament connection just described in the example requires that two different types be carried as spares.

The current rating of the rectifier tubes can sometimes be increased by operating the rectifier filaments in *quadrature*. This is the term applied to rectifier operation when the filament voltage is 60 to 120° out-of-phase with the voltage applied to the rectifier tube plate. Transformer connections for quadrature operation of the rectifier filaments are illustrated

TABLE 8-1. TYPICAL POWER-RECTIFIER CIRCUIT CONNECTIONS AND CIRCUIT DATA

Type of circuit	Rectifier	Single-phase full-wave	Single-phase full-wave (bridge)	Three-phase half-wave	Three-phase half-wave	Six-phase half-wave	Six-phase half-wave	Six-phase half-wave π fork	Six-phase (double-three phase) half-wave	Three-phase full-wave	Three-phase full-wave
	Transformer	Single-phase center-tap	Single-phase (bridge)	Delta-wye	Delta-zig-zag	Delta-star	Delta six phase π fork	Delta W double wye with balanced coil	Delta-wye	Delta-delta	
Circuit	Secondary										
	Primary										
	Number of phases of supply	1	1	3	3	3	3	3	3	3	3
	Number of rectifiers	2	4	3	3	3	6	6	6	6	6
	K_f factor	0.12	0.12	0.03	0.03	0.0034	0.0034	0.0034	0.0034	0.0034	0.0034
	Ripple voltage	0.48	0.48	0.18	0.18	0.042	0.042	0.042	0.042	0.042	0.042
	Ripple frequency	2f	2f	3f	3f	6f	6f	6f	6f	6f	6f
	Line voltage	1.11	1.11	0.855	0.855	0.740	0.740	0.740	0.855	0.855	0.855
	Line current	1	1	0.816	0.816	0.816	0.816	0.816	0.707	0.707	0.816
	Line power factor ϕ	0.90	0.90	0.826	0.826	0.955	0.955	0.955	0.955	0.955	0.955
Transformer primary: volts per leg	1.11	1.11	0.855	0.855	0.740	0.740	0.740	0.855	0.855	0.855	
amp per leg	1	1	0.471	0.471	0.577	0.577	0.577	0.408	0.408	0.471	
kva	1.11	1.11	1.21	1.21	1.28	1.28	1.28	1.05	1.05	1.05	
Transformer average	1.34	1.11	1.35	1.46	1.55	1.55	1.42	1.26	1.05	1.05	
kva	1.11A	1.11	0.855	0.855	0.740A	0.740A	0.428A	0.855A	0.428	0.428	
Transformer secondary: volts per leg	0.707	1	0.577	0.577	0.408	0.408	0.577B (0.408C)	0.289	0.816	0.816	
amp per leg	1.57	1.11	1.48	1.71	1.81	1.81	1.79	1.48	1.05	1.05	
kva	3.14	1.57	2.09	2.09	2.09	2.09	2.09	2.42	1.05	1.05	
Peak inverse voltage per rectifier	1	1	1	1	1	1	1	0.5	1	1	
Peak current per rectifier	0.5	0.5	0.333	0.333	0.167	0.167	0.167	0.167	0.333	0.333	
Average current per rectifier	0.5	0.5	0.333	0.333	0.167	0.167	0.167	0.167	0.333	0.333	

* These circuit factors are equally applicable to electron-tube or metallic-plate rectifiers.

† (Line power factor) = (d-c output watts) / (line volt-amperes).

Note:

Unless otherwise stated, factors shown express the ratio of the rms value of the circuit quantities designated to the average output values of the rectifier.

Factors are based on a sine-wave voltage input, infinite-inductance choke, and no transformer or rectifier losses.

in Fig. 8-9. The filament transformers for the rectifiers are supplied from each of the three phases. These are different from those of the plate transformer supplying that particular rectifier tube.

It is apparent from the basic rectifier circuits that the filament transformers for the rectifiers must be capable of withstanding the voltage involved in that particular rectifier circuit.

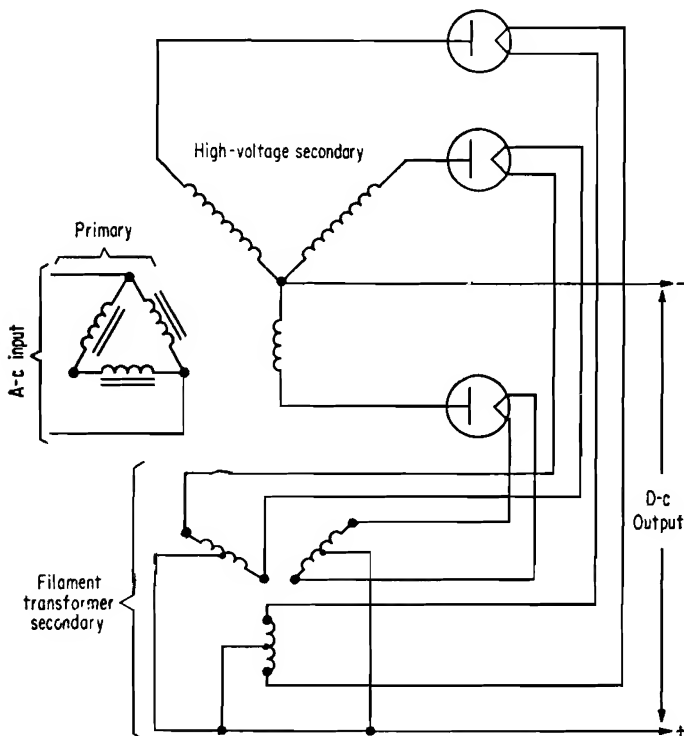


FIG. 8-9. Typical filament connections for quadrature operation of high-voltage rectifier.

The choice of the rectifier circuit used in a particular application depends on the a-c power source available and the economic considerations based on the cost of the components used in the various possible circuits. For example, a single-phase full-wave bridge circuit when compared to a three-phase full-wave circuit would require a higher-capacity filter network since the ripple frequency is three times lower. However, the former rectifier circuit also has the advantage of requiring the use of only four rectifiers, while the latter type of rectifier requires six. This advantage must then be weighed against the fact that the single-phase bridge rectifier requires a rectifier with a higher-peak

inverse-voltage rating than the same voltage supply operating as a full-wave three-phase supply.

In general, the polyphase rectifier circuits are preferred for use in power supplies with an output rating over 1 kva, while single-phase systems are better suited to power outputs less than 1 kva.

8-4. Rectifiers. Transmitter power supplies utilize both the thermionic and the semiconductor types of rectifiers. The specific rectifier choice is determined by the application, the service conditions, and the cost. The thermionic rectifiers include the gas-filled and the high-vacuum types, while the semiconductor types include the copper oxide, selenium, germanium, and silicon rectifiers.

The rectifier characteristics which govern the electrical suitability of any rectifier for a given application are the peak inverse-voltage rating of the rectifier, the average d-c plate-current rating, and the peak plate-current rating. The peak inverse-voltage rating is the peak voltage that the tube can safely and continuously withstand in the nonconduct-

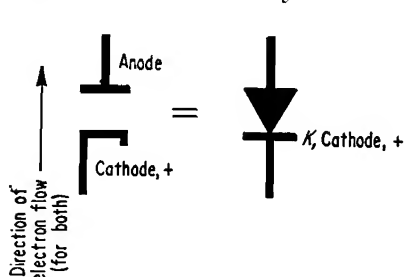


FIG. 8-10. Equivalent symbols and direction of electron flow for thermionic and semiconductor rectifiers.

ing direction, i.e., when the plate is negative and the cathode is positive. The d-c plate-current rating is the value of the average d-c current that can be continuously passed through the tube. This is not always the value of the d-c plate current in the load. It will vary with the various rectifier circuits shown in Table 8-1. The peak plate-current rating through the rectifier tube is the maximum plate current that the tube can safely

pass. This current is largely a function of the filter network following the rectifier.

While the function of the thermionic rectifiers is the same as the semiconductor types, the symbols used in electronic circuitry are different. The electrical equivalents of the two general types are illustrated in Fig. 8-10.

Gas-filled Rectifiers. General. This type of rectifier utilizes the electron bombardment of gas atoms to produce positive gas ions. These cancel the normal negative space charge, and the result is a tube with a very low internal voltage drop. The type of gas used in the rectifier determines the voltage drop. Typical gases used include argon, hydrogen, helium, mercury, and xenon. The latter two are most commonly used.

Mercury-vapor Rectifiers. This type of rectifier uses a small pool of mercury which is vaporized by the filament to form a mercury vapor.

For proper operation, the mercury vapor must be operated within the temperature limits stipulated by the tube manufacturer. The proper operating temperature is determined by the condensed-mercury temperature at the base of the bulb. This temperature can be measured with a thermometer or thermocouple attached to the base of the rectifier with putty. A high condensed-mercury temperature will decrease the voltage at which the tube begins to conduct. This will increase the tube filament life, but it will also reduce the peak inverse voltage that the tube will stand. Conversely, a low condensed-mercury temperature raises the voltage at which the tube will begin to conduct, and it will also shorten the filament life.

A few mechanical precautions should be observed when using mercury-vapor rectifiers. Mercury-vapor rectifiers are usually designed to be mounted in a vertical position. This should be observed unless otherwise stated by the tube manufacturer. Second, the life of mercury-vapor rectifiers can be severely shortened by exposing them to concentrated r-f or magnetic fields. Therefore, shielding the rectifiers may be necessary.

Xenon Rectifiers. This type of rectifier has all the advantages of the gas-filled rectifier without the operating-temperature limitations of the mercury-vapor rectifier. Typical xenon rectifiers can operate from -75 to $+90^{\circ}\text{C}$. The typical voltage drop of the xenon-filled rectifier is only 10 volts. Furthermore, this type of rectifier can usually be mounted in any position.

Vacuum Rectifiers. This type of rectifier depends only on thermionic emission for its unidirectional current-carrying characteristics. This type of rectifier is characterized by a high internal voltage drop. It finds wide use in relatively low-voltage plate supplies, i.e., those not over 500 volts, at 250 ma. At these levels, rectifier heating and efficiency are not usually of prime concern. The high-vacuum rectifier is also capable of very high-peak inverse-voltage ratings. This makes its use practical in power supplies where the extremely high-peak inverse-voltage ratings required make the use of other types of rectifiers impracticable. In this latter case, high-vacuum rectifiers can be used in rectifier circuits producing upwards of 20,000 volts at 2 amp. At these levels the internal drop is of little consequence. The resultant plate dissipation of the rectifiers can be adequately cared for by either forced- or natural-air circulation.

Arckback in Thermionic Rectifiers. An *arckback* is defined as a failure in the rectifying action which results in the flow of a principal electron stream in the reverse direction because of the formation of a cathode spot on the anode. The arckback then occurs when the polarity of the rectifier reverses and the plate is negative with respect to the cathode.

Arcback is a common fault in high-voltage rectifiers, and, if the protective devices are adequate, it will result in no damage to the rectifiers. In such cases, the service can be immediately restored. Arcbacks occur in a random fashion without warning or previous indication. While arcbacks are an unfortunate by-product of rectifier operation, they do not necessarily indicate a poor rectifier if the frequency of occurrence is reasonable. What is reasonable depends on the rectifier service and the operating conditions and is best determined from experience with each type of rectifier in a particular piece of equipment. Poor vacuum

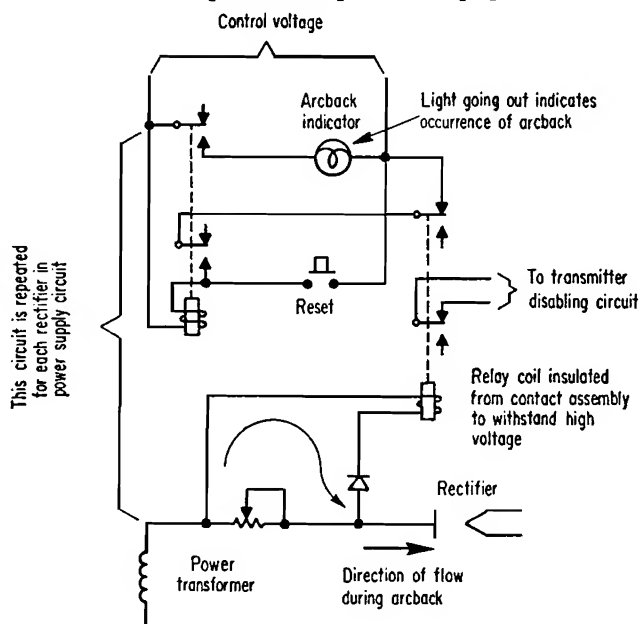


FIG. 8-11. Typical arcback circuit and indicator circuit.

and excessively high operating temperatures will increase the chances of rectifier arcback.

Rectifier Protection against Arcback. Arcback will not damage a rectifier if the supply can be deenergized fast enough to prevent the arcback current from reaching dangerous proportions. The most satisfactory method is based on detecting the reverse current flow through the rectifier rather than relying solely on the power-supply alternating current over current breakers. The directional-current-sensing device usually makes use of a rectifier-and-relay combination in the plate lead of each rectifier. This directional-current relay is arranged to disconnect the a-c line from the power supply by opening a contactor or undervoltage-release solenoid of a circuit breaker. A typical method is shown in Fig. 8-11.

Semiconductor Rectifiers. The semiconductor rectifier makes use of the unidirectional characteristics of copper oxide, selenium, germanium, and silicon junctions to achieve rectification. The mechanics of the rectification process are complicated and beyond the scope of this book (1).

The copper oxide rectifier has a low maximum voltage rating per cell which makes its use for high-voltage applications cumbersome and uneconomical. Furthermore, its use in low-voltage applications is limited by the relatively large size of the unit compared to the other rectifier types available today. For this reason copper oxide rectifiers find little use in new equipment.

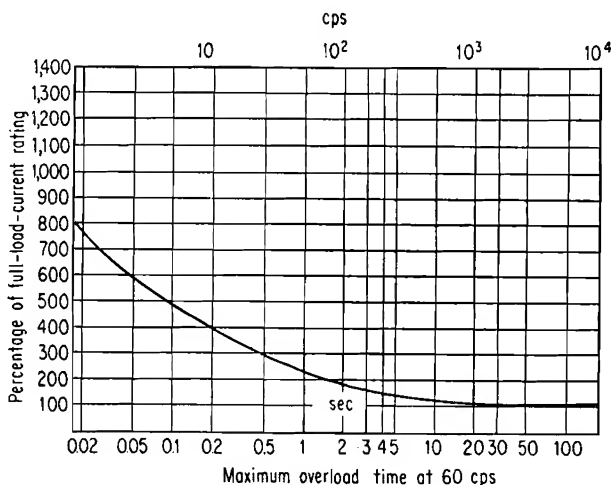


FIG. 8-12. Typical overload curve for germanium and silicon rectifiers.

Selenium rectifiers have a higher a-c voltage rating of approximately 26 volts per cell. However for high-voltage applications a large number of cells would still be required. Selenium rectifiers also suffer from "aging"; i.e., both the forward and reverse resistance increase with use. Thus, to maintain a constant d-c output voltage, means must be provided for increasing the a-c input voltage to compensate for the increased voltage drop in the rectifier.

Germanium and silicon rectifiers have a-c voltage ratings of up to 1,000 volts per cell, thus making the use of this rectifier quite practical in high-voltage circuits. However, because of the small rectifying junction area, any overloads will quickly overheat the cell and possibly destroy it. Thus this type of rectifier requires special consideration to protect it against overloads. Figure 8-12 gives a curve which is typical of the overload characteristics of germanium and silicon rectifiers.

Overload Protection of Semiconductor Rectifiers. Copper oxide and selenium rectifiers have the thermal capacity in the rectifying junctions to enable satisfactory protection against overloads to be achieved with ordinary properly rated fuses and circuit breakers. For example, the overcurrent protective circuits need only deenergize a power supply using selenium rectifiers in 10 cycles (based on 60 cps) or in 0.166 sec. Circuit breakers and other protective devices are usually rated in cycles in this manner. The overcurrent speed for copper oxide rectifiers is even slower, since it has a greater thermal capacity attributable to its larger size. However, both germanium and silicon rectifiers require protection which will act in one-half cycle or less in order to prevent damage to the rectifier. Special high-speed fuses (such as Chase Shawmutt's Amp-Trap or Bussman's Silver Strand) are often used for this purpose. Furthermore, because of the low internal impedance of this type of rectifier, higher short-circuit currents can be expected. Some derating of the rectifier is often used to permit a longer overload time. However using a rectifier much larger than is necessary to handle the normal current flow will increase the cost of the rectifier considerably.

Transient Protection of Semiconductor Rectifiers. Because of the existing inductances and capacitances in a power supply, large voltage peaks or transients are often built up when a power supply is turned on or off. These transient voltages may be in excess of the voltage rating of the rectifier units and therefore may result in a damaged rectifier with short-circuited cells. These transient voltages are not distributed evenly over a series string of rectifiers. The highest transient voltages will appear across the rectifier furthest from ground. Thus one cell at a time may be short-circuited owing to the transient overvoltage. This will eventually result in failure of the entire remaining rectifier cells in the series string.

An analysis of the transient conditions existing in typical power supplies shows that this is not normally a problem in rectifiers utilizing large-area rectifying junctions such as in the typical selenium or copper oxide cell. However, the rectifying cell area of the typical germanium or silicon rectifier is very small, and transient protection must be considered. One obvious solution is to simply ensure that the voltage rating of the rectifier being used is higher than the expected or measured transient voltage. However, this is usually the most expensive solution since the transient voltages are commonly four or five times the normal operating voltages.

A more economical solution is to employ nonlinear resistors in the circuit to protect the rectifiers from the voltage peaks. The nonlinear resistor is actually a resistor that is voltage-sensitive. In this case the higher the voltage across the ends of the resistor, the lower will be its

resistance. The nonlinear resistors commonly used in this application are distributed under the trade name of Thyrite and manufactured by the General Electric Company. The connection of the Thyrite protective resistors is shown in Fig. 8-13. Multiphase half-wave rectifiers

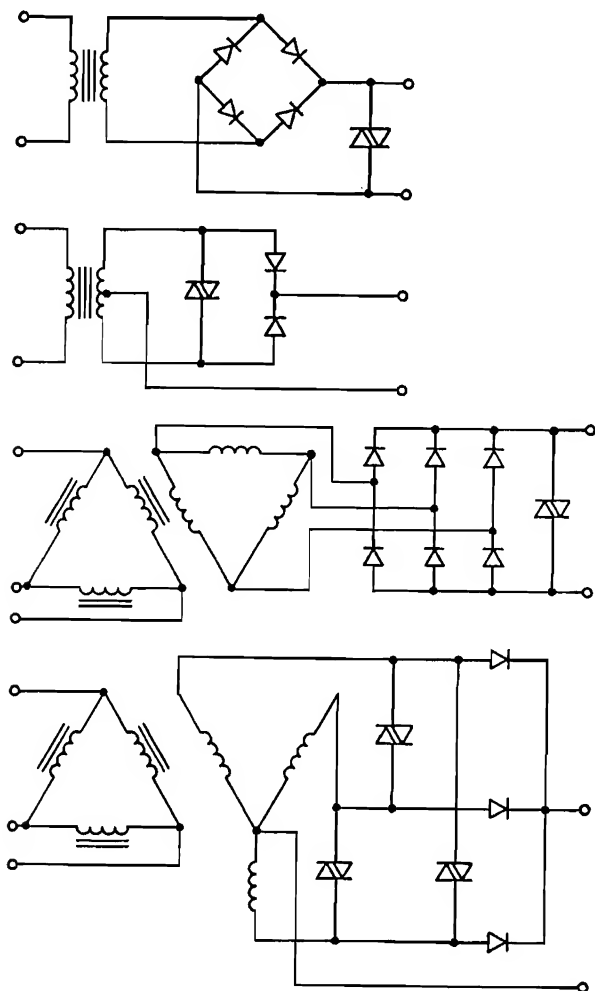


FIG. 8-13. Application of nonlinear resistors (Thyrite) for transient protection of germanium and silicon rectifiers.

require the use of as many protective Thyrite resistors as there are phases, while the full-wave applications require the use of only one (2).

Capacitors can also be used for transient protection. A $0.01\text{-}\mu\text{f}$ capacitor connected across each rectifier cell will usually serve to hold the peaks down to tolerable levels. In rectifier assemblies with many

cells in series, a 0.001- μ f capacitor can be connected across each series of six cells for transient protection (3).

Current Ratings of Semiconductor Rectifiers. The current rating of a semiconductor rectifier depends on the temperature of the rectifying junction. Since measuring the junction temperature is not normally too convenient for a rectifier user, a manufacturer will rate a rectifier on the basis of normal free air with convection cooling with a maximum ambient operating temperature. In this way, the rectifier junction

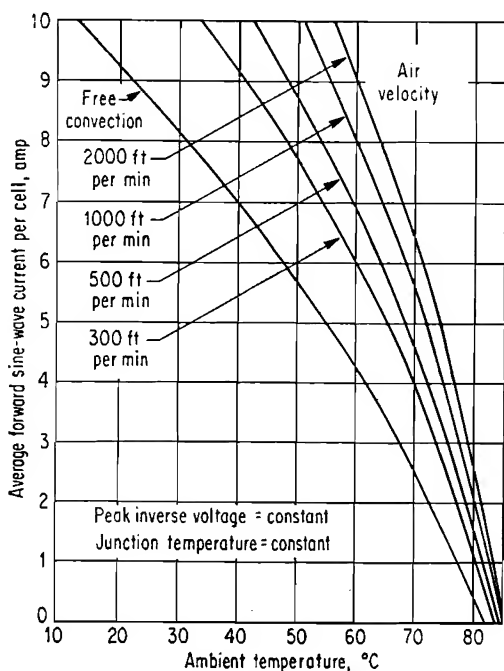


FIG. 8-14. The effect of cooling on the current-carrying capacity of a typical germanium rectifier for different ambient temperatures.

temperature will not exceed the specified maximum. However, rectifier ratings can be raised through the use of forced-air cooling or by immersing the rectifier in oil or some other coolant. Figure 8-14 illustrates the increased current rating obtained through the use of forced-air cooling.

8-5. Power-supply Filter Circuits. There are two general power-supply filter configurations, namely, the capacitor input and the choke input types. These are shown in Fig. 8-15. The use of a capacitor input filter results in reduced rectifier current-carrying capacity as a result of the high current peaks through the rectifier during the charging time of the input capacitor. The capacitor input filter is also characterized by poor voltage regulation, since the output voltage is determined

by the load resistance and the value of the input capacitor. Because of the disadvantage of this filter system, it is not usually used in transmitter power supplies. The calculation of the performance of the capacitor input filter is very difficult, and in many cases is best determined experimentally. However, where all the parameters of the circuit are known, the performance can be calculated (4).

The Choke Input Filter. This type of filter is characterized by its superior voltage-regulation characteristic as compared to the capacitor input filter. A comparison of the voltage-regulation characteristics for these two filter systems is made in Fig. 8-16. The choke used in this

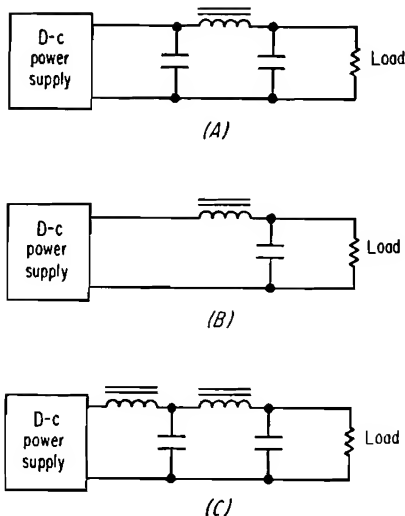


FIG. 8-15. Typical power-supply filters. (a) Capacitor input filter; (b) choke input filter; (c) two-section filter.

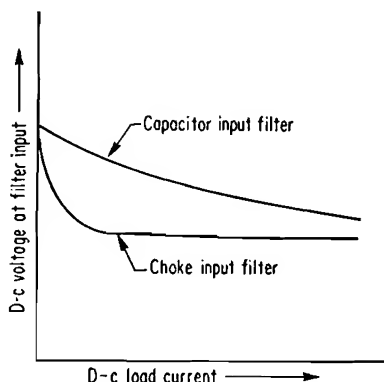


FIG. 8-16. Comparison of d-c output voltage with changes in load current for typical capacitor and choke input filters.

type of filter must possess sufficient inductance to ensure that the rectifiers are continuously supplying power to the filter. This is contrasted to the capacitor input filter, in which the power is delivered to the filter for only a portion of each power cycle.

For economic reasons the choke input filter is the most commonly used type of filter in transmitter circuitry. Use of this type of filter results in a higher utilization factor (see Sec. 8-6) and lower peak rectifier currents. This will result in a smaller and more economical transformer as well as a lower rating or at least a longer life for the rectifier.

In general, the higher the gain of the stages supplied by the power supply, the greater will be the smoothing required of the a-c ripple from the rectifier. This requirement of the power supply is governed by the nature of the load and is best expressed as a ratio called the *ripple factor*.

$$\text{Ripple factor} = r = \frac{E_{ac}}{E_{dc}} \quad (8-2)$$

where E_{ac} = a-c rms ripple voltage at filter output

E_{dc} = d-c voltage at filter output

Typically, a high-gain audio stage might require a ripple factor in the order of 0.0005 to 0.005; and a high-voltage r-f power amplifier power supply might be required to have a ripple factor of 0.01 to 0.1.

If the value of the ripple factor has been determined (from the load considerations) then the smoothing factor of the power-supply filter can be determined. The smoothing factor is expressed by the ratio

$$\text{Smoothing factor} = \alpha = \frac{E_{in}}{E_{out}} = \frac{E_r}{r} \quad (8-3)$$

where E_{in} = a-c ripple voltage into filter

E_{out} = a-c ripple voltage at output of filter

E_r = ripple voltage expressed as a ratio of ripple voltage to d-c output voltage. (This is determined by the rectifier circuit; see Table 8-1 for the value.)

r = ripple factor determined from Eq. (8-2)

The smoothing factor determined in Eq. (8-3) can be substituted in the Eq. (8-4) to determine the value of the LC product required by the filter.

$$LC = \frac{\alpha + 1}{(2\pi f_r)^2} \quad (8-4)$$

where α = smoothing factor determined in Eq. (8-3)

f_r = ripple frequency at input to filter, cps (see Table 8-1 to determine the ripple frequency.)

The filter choke and filter capacitor can be any value providing (1) the product of the two exceeds the value determined in Eq. (8-4) and (2) the filter-choke inductance exceeds the minimum value required to maintain a continuous current through the input choke. This value which includes a 25 per cent safety factor is given by the equation

$$L_{min} = \frac{KR_L}{f_r} \quad (8-5)$$

where L_{min} = minimum inductance required by filter choke, henrys

K = factor determined from Table 8-1

R_L = maximum value of load resistance, ohms

f_r = ripple frequency at input to filter, cps (See Table 8-1 for value.)

In many cases, the value determined for the filter is physically unrealizable or is otherwise uneconomical. In many of these cases, it may be desirable to use two cascaded sections of filtering as shown in Fig. 8-15c.

The values of the filter choke and capacitance can be determined by the equation

$$LC = \frac{0.0254 (\sqrt[n]{\alpha} + 1)}{f_r^2} \quad (8-6)$$

where LC = product of inductance, henrys, and capacitance, farads, for each section of filter

n = number of filter sections

α = smoothing factor for total filter determined in Eq. (8-3)

f_r = ripple frequency, cps

Even though a multisection filter reduces the values of the filter capacitor, care must be used to ensure that the application does not require a larger value for purposes other than filtering. For example, in TV r-f power amplifiers, where the plate-current demand is at a video rate, the power-supply output filter must serve as a low impedance to all the video frequencies involved in the amplifier operation. This will also apply to transmitters which are keyed at a fast rate, where the current demand on the power supply is varying.

Although there is considerable latitude in values of the components used in power-supply filters, a calculation should be made after the values have been calculated by the equations above to determine that the values chosen are not resonant at the lowest ripple frequency that would be encountered in the filter. To include all eventualities such as rectifier unbalance or failure, this would be the input power frequency, even in cases where the output ripple frequency is a multiple of the input frequency. A series resonance that might take place between the inductance and capacitance would make the ripple much larger than at the input. The values of the inductance and capacitance determined by the above means can be inserted in the following equation to determine the resonant frequency. As pointed out above, this should preferably be lower than the input power-line frequency.

$$F = \frac{1}{2\pi \sqrt{LC}} \quad (8-7)$$

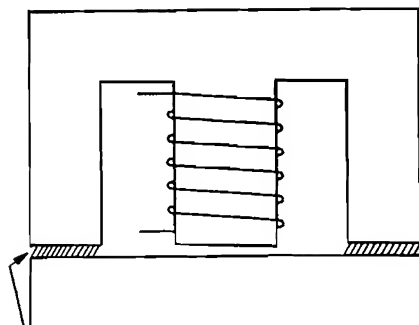
where F = resonant frequency of filter, cps

L = filter-choke inductance, henrys

C = filter capacity, farads

The Swinging Choke. The inductance of a swinging choke depends on the current flowing through the choke and, within limits, is inversely proportional to the current flowing through the choke. From the power-supply filter equation (8-5) given previously, it can be seen that the value of the minimum inductance increases as the load resistance increases.

Therefore, for low currents, a higher value of minimum inductance is required. For varying loads, such as a class B amplifier, some economy



Nonmagnetic spacer, the wider the gap, the less the inductance and higher the saturating current through the choke

FIG. 8-17. A simplified swinging choke. The wider the gap, the less inductance, and higher the saturating current through the choke.

can be effected in the choke since a swinging choke will usually result in a smaller and hence lower-priced choke. A simplified swinging choke is shown in Fig. 8-17.

The Bleeder Resistance. For a given choke inductance, the load resistance must never increase beyond the value specified in Eq. (8-5). The load resistance and the bleeder resistance together form the load and the resistance specified in this equation. For varying loads, the value of the bleeder resistance is calculated to ensure that the value of the total load resistance is never higher than that

calculated in Eq. (8-5). Specifically, the maximum bleeder resistance can be calculated as

$$R_b = \frac{R_m R_L}{R_L - R_m} \quad (8-8)$$

where R_b = maximum bleeder resistance to satisfy minimum inductance for choke input filter

R_L = maximum load resistance excluding bleeder resistance

R_m = maximum total load resistance including bleeder used in Eq. (8-5)

8-6. Transformers and Chokes. General. Transformers and chokes that operate at voltages below 7,500 volts are usually of the dry type, while above this voltage oil-filled types are generally used. The use of oil in a transformer reduces the necessary insulation required in the unit. For example, a 10-in. creepage distance is necessary in air to withstand a 1-min breakdown test of 60 kv. In oil, only a 2-in. distance is necessary to withstand this same voltage. Thus the use of oil will result in a smaller and more compact transformer at the higher voltages.

Oil-filled transformers are often conveniently combined with other transformers and chokes in one tank. This technique results in a further reduction in over-all size. A typical unit of this type might contain a three-phase rectifier filament transformer, a three-phase high-voltage plate transformer, and a filter choke. If semiconductor rectifiers were

used, these could also be included. While this increases the current rating of the rectifiers, it also presents a maintenance problem.

Oil-filled transformers and chokes can use either standard mineral transformer oil which is flammable, or a synthetic nonflammable insulating liquid known as *Askarel*. Indoor installations of transformers using flammable transformer oil require the use of a separate transformer vault in accordance with the National Electric Code. These vaults should be fireproof, and are arranged to drain the oil safely outside in case of a leak.

Askarel is also used under other trade names, such as Aroclor, Dykanol, Chlorextol, Inerteen, and Pyranol. In any case, it should be kept free from contamination with water, and transformer tanks should be sealed to prevent evaporation of the liquid.

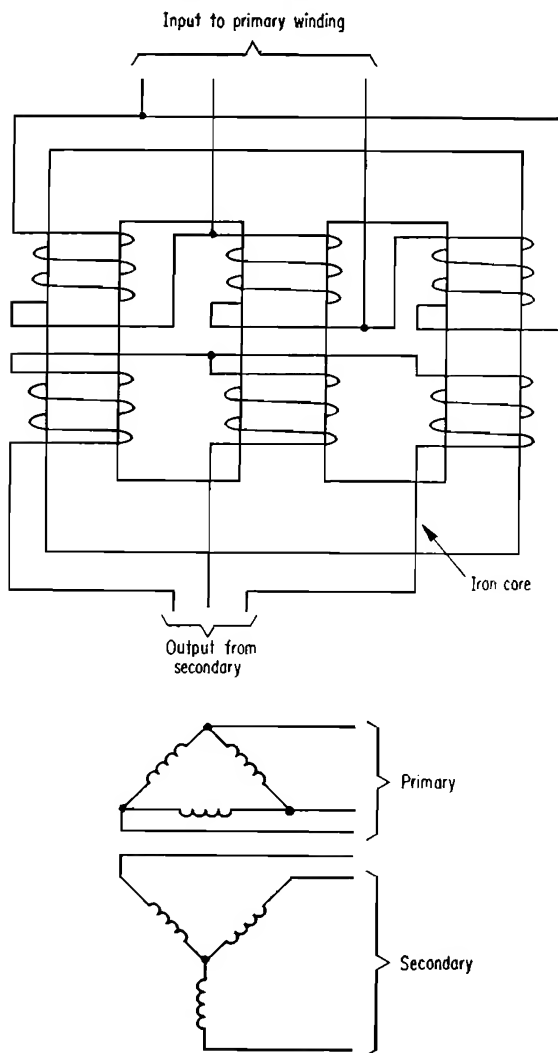
Transformer Utilization. The various rectifier circuits will result in different waveshapes of current flowing through the primary and secondary windings. Since these current waveforms are not sinusoidal, the resultant heating of the transformer will be greater for these currents than for the equivalent a-c power delivery to a resistance load.

The ratio of the d-c power output from a power supply to the kilovolt-ampere capacity which results in the same heating of the winding is the transformer utilization factor. This is given in reciprocal form in Table 8-1 for the various rectifier circuits. It will be noted that this factor varies with the rectifier circuits and is an indication of the comparative size and hence relative cost of the equivalent transformers used in the different circuits.

The transformers selected for the three-phase power-supply circuits shown in Table 8-1 can be individual transformers, or they can be combined on one core into one unit. However, for the three-phase half-wave circuits the use of individual units will result in the d-c saturation of the core, unless the transformer iron is adequate to take care of the magnetization introduced by the d-c current flow through the secondary. In these circuits, a smaller transformer will result if the entire transformer is wound on one core. The d-c magnetization that resulted from the use of individual units is canceled out. A typical three-phase transformer wound on a single core is shown in Fig. 8-18.

Choke Current Interruptions. A large current change occurs in a choke whenever it is turned on or off and particularly whenever a fault occurs in the power-supply load. This sudden surge of current will induce a large voltage in the choke, which, if not suitably damped, may result in permanent damage to the choke. The induced voltage, which is the result of the current change is expressed mathematically as

$$e = -L \frac{\Delta i}{\Delta t} \quad (8-9)$$



Typical three-phase transformer construction with electrical-circuit equivalent

FIG. 8-18. Typical three-phase transformer construction corresponding electrical schematic.

where e = induced voltage

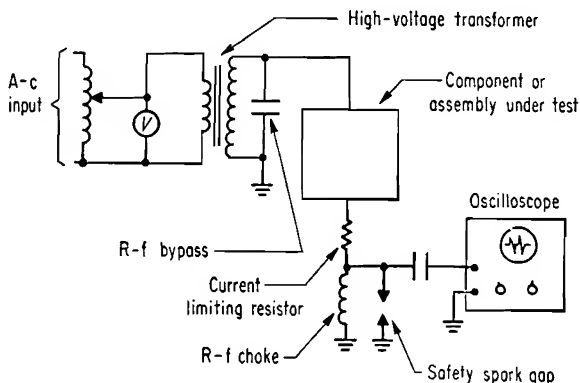
L = inductance of choke, henrys

$\Delta i / \Delta t$ = rate of change of current, amp, with respect to time, sec

Equation (8-9) illustrates the fact that the faster the current changes through the choke the higher will be the voltage induced in the choke. The simplest means of protecting the choke against these extremely high-

voltage surges is by means of ball spark gaps placed directly across the terminals of the choke, or through the use of hermetically sealed spark gaps specially designed for this purpose.

8-7. Corona. Corona is the ionization of the dielectric material surrounding a conductor. This is the inevitable result of overstressing the dielectric material (usually air) by high voltage. The presence of corona can be detected by the physical senses; i.e., it can be heard as a hissing or frying sound; or it can be seen as a dark pale glow, or detected by the characteristic odor of ozone. Ozone is a highly reactive oxidizing gas which attacks the insulation. Corona also produces nitric acid and nitrous acid in the presence of moisture. The latter acid will readily attack silver, lead, and copper. These factors, combined with the heat



Schematic for detecting corona with an oscilloscope

FIG. 8-19. Schematic for detecting corona with an oscilloscope.

generated by the corona, result in a rapid deterioration of the insulation and metals involved. Thus it is apparent that corona should be eliminated wherever it forms.

There are four general methods that can be used to eliminate corona:

1. Increase the spacing between components and conductors of widely different voltages.
2. Improve the shape of the components, terminals, or conductors. By elimination of sharp points, edges and corners on components at high voltages, the voltage stresses will be reduced, and the voltage at which corona begins will be increased markedly.
3. Provide corona shields; these not only surround any sharp points, edges, and corners but also redistribute and reduce the voltage stresses.
4. Eliminate the air gaps present by use of epoxy or other molding materials or by immersing the component in oil.

Proper maintenance procedures will help to minimize the formation of corona. All insulating surfaces should be kept dry and clean. Lower levels of corona can be detected by the circuit illustrated in Fig. 8-19. This method makes use of an oscilloscope as a corona detector. Corona is seen on the oscilloscope screen as spurious voltage pulses.

The actual level of corona can be measured by the circuit illustrated in Fig. 8-20. This method is based on the standard NEMA corona test. The signal strength noted on the receiver due to corona is compared to the calibrated output of a standard signal generator. Thus the corona signal level in microvolts is determined. With the applied voltage 5 per cent higher than normal, the corona level for transformers with up to 8.6 kv output should be less than 1,000 μv . For transformers with

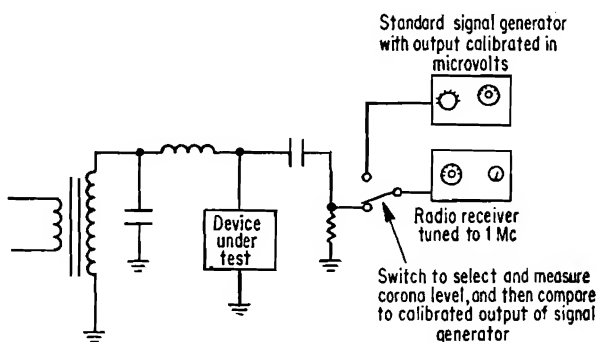


FIG. 8-20. Schematic for measuring corona level.

outputs between 8.61 and 15 kv, the maximum permissible corona level is 2,500 μv .

8-8. Filament Starting. As transmitting tubes increase in power rating, the filament requires more power to supply the increased emission demand. The heavy current requirements of the larger power tubes, together with the large change in the cold to hot resistance of the filament requires extra consideration in supplying the filament power to these tubes. The typical cold resistance of a tube is one-tenth that of the hot resistance. This figure will vary according to the type of filament and its operating temperature. However, using this figure as typical, the starting current of a tube would be 10 times the current at the final operating currents. This large current sets up magnetic and physical stresses within the tube structure that may destroy the tube if not controlled. For this reason, the filament current is usually required to be limited to a safe value specified by the tube manufacturer. Some methods automatically limiting the filament current to the maximum value are illustrated in Fig. 8-21. This can be done by using a variable autotransformer, which is initially adjusted to zero-volt output and then

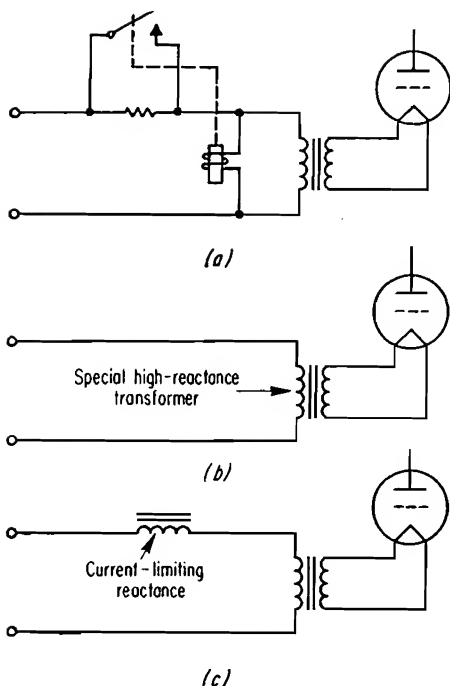


FIG. 8-21. Typical methods for limiting filament starting current.

increased slowly such that the maximum filament current is not exceeded. This is normally not too satisfactory since it requires that the operator remember this operation each time the transmitter is turned on and off.

The method illustrated in Fig. 8-21a uses a current-limiting resistance in series with the filament transformer primary. This automatically cuts out when the filament current drops to a predetermined value. The method illustrated in Fig. 8-21b makes use of a high-reactance or current-limiting transformer to hold the current below the maximum specified. The general construction of this type of transformer is illustrated in Fig. 8-22. This transformer has a magnetic shunt between the primary and secondary windings. At no load, these shunts carry very little flux. However, as the load increases, the second-

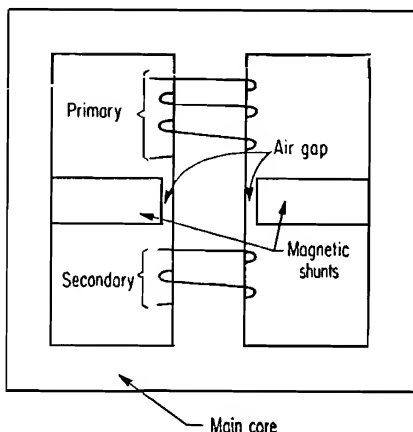


FIG. 8-22. Basic construction of a high-reactance transformer.

ary ampere-turns will increase the flux forced through the magnetic shunt. This decreases the magnetic coupling between the primary and secondary windings and hence reduces the output voltage from the secondary. The

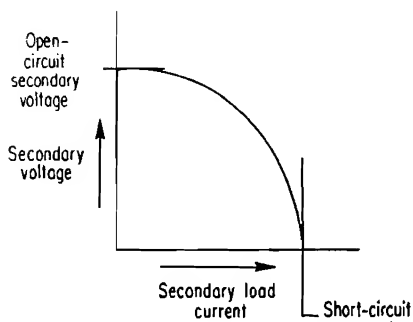


FIG. 8-23. Secondary output voltage vs. load current for a high-reactance transformer for use in filament starting.

general characteristic curve for a high-reactance transformer of this type is shown in Fig. 8-23. Another method of limiting the filament current is shown in Fig. 8-21c. This method uses the high reactance of a choke (or reactor) in series with the primary of a filament transformer. If the load resistance is low as it is when the filament is initially turned on, then the voltage drop across the series reactor will be high, and the drop across the transformer will be low. As the filament resistance increases, the voltage drop across the reactor will decrease, and the voltage drop across the transformer primary will increase until the filament resistance reaches its operating temperature and resistance.

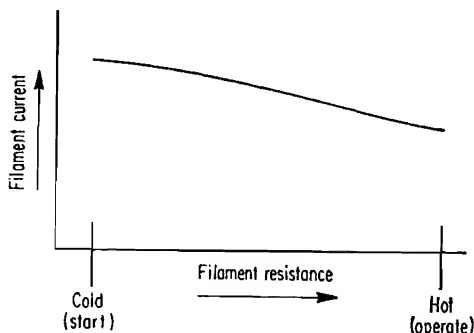
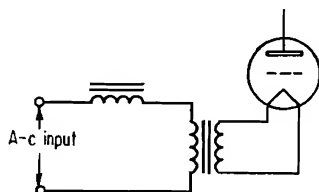


FIG. 8-24. Filament current-limiting characteristic using a series reactance.

In this manner, a constant-current characteristic can be approached—a typical characteristic curve. The operating characteristics of this method are illustrated in Fig. 8-24.

Multiphase Filament Supplies. Many of the large transmitting tubes employ multistrand filaments designed to be operated from the different phases of a polyphase system. This results in a substantial hum reduction, since some part of the filament is always receiving power. The hum frequency resulting from multiphase operation of the filament is also increased. The filament operating from a 60-cps source normally results in a 120-cps a-m hum in the amplifier output. A filament operated from a 60-cps, three-phase source will result in a 360-cps a-m hum in the carrier, while a six-phase source will produce a 720-cps a-m hum in the carrier. Filaments operated in this manner are shown in Fig. 8-25.

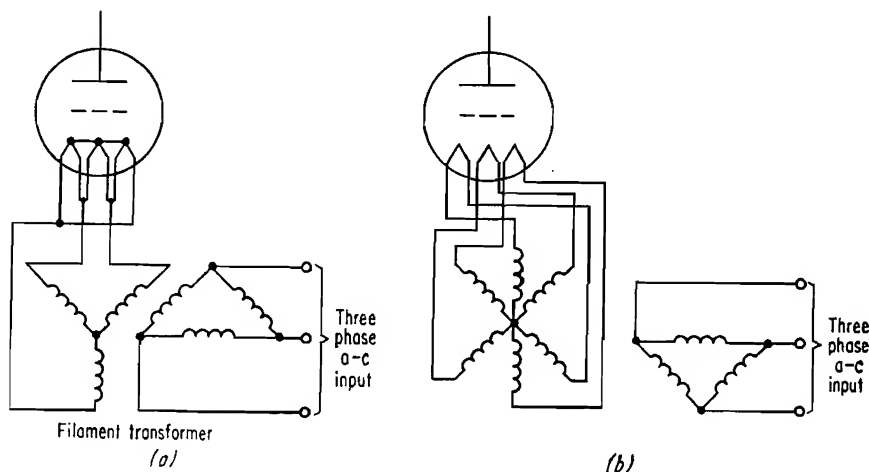


FIG. 8-25. Typical multiphase connections for tube filaments. (a) Three-phase excitation of tube filament; (b) six-phase excitation of tube filament.

Some transmitting tubes are designed to be operated from a two-phase source. Since the majority of equipment is operated from a three-phase source, the use of the Scott connection can be used to produce the desired two-phase excitation. A typical circuit and the vector diagram of the voltages and the phase relationships in this circuit is shown in Fig. 8-26.

D-C Filament Supplies. In many applications, the hum introduced by operating the filaments from an a-c source is objectionable. In these cases operating the filament from a d-c source becomes the most convenient method of eliminating this hum. The design of these power supplies would follow the same procedure previously described in Sec. 8-3. To limit the filament starting current, the means previously described should still be used. Usually, the germanium and silicon rectifiers are used for this application because of their high current capability and their low internal voltage drop.

8-9. D-C Regulators. Many transmitter circuits require close voltage control of the applied d-c voltages. There are many methods of accomplishing this control. The exact method depends on the degree and preciseness of the voltage control. Most commonly used regulators make use of a series or shunt passage tube. These tubes are controlled by the regulator amplifier which detects the voltage change, compares it to a reference voltage, and biases the d-c control tube to compensate

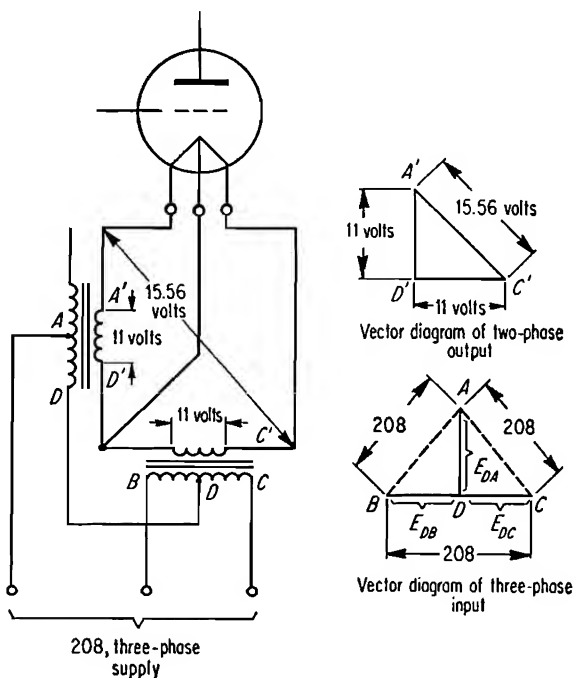


FIG. 8-26. Scott-tap filament connection for obtaining two-phase filament power from three-phase power line.

for the change. A block diagram illustrating the functions of a d-c voltage regulator is shown in Fig. 8-27.

The shunt tube used as a d-c regulator tube must be capable of dissipating the total change in the load current at the regulated output voltage from the power supply. The series tube must be capable of dissipating the difference between the input voltage to the d-c regulator and the output voltage from the regulator at the maximum current flow to the load. Thus for large power supplies, the power-handling capacities of these tubes will be quite large. In many cases, tubes are operated in parallel to obtain the required dissipation capability. If tubes are

operated in parallel, it is necessary to use current-equalizing resistors in the cathode leads of each tube to equalize the current division among the tubes owing to the slightly different characteristics of each tube.

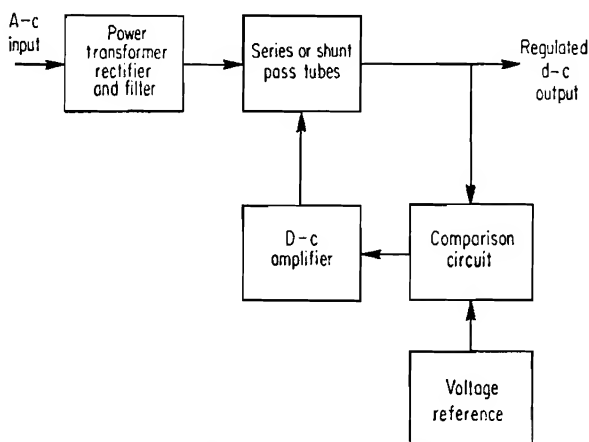


FIG. 8-27. Block diagram of a voltage-regulated power supply.

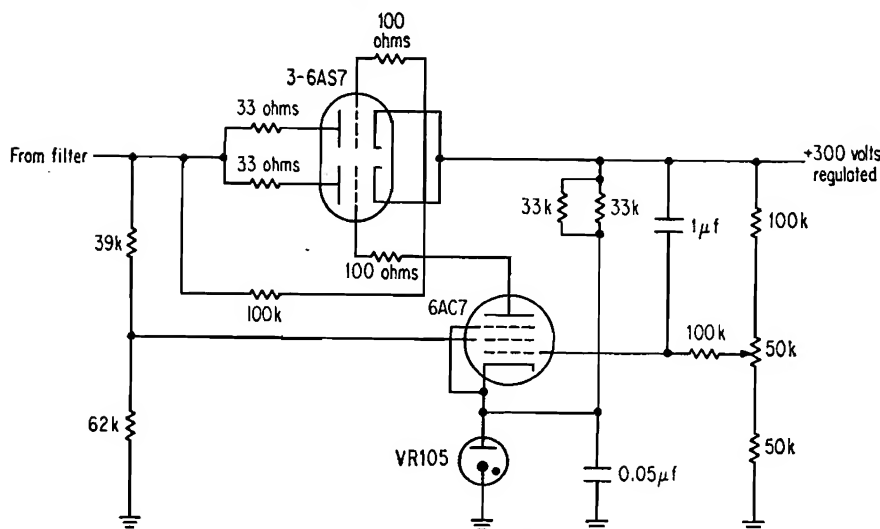


FIG. 8-28. Typical series voltage regulator.

The complexity of the regulator amplifier depends on the desired regulation capability of the regulator. If regulation no better than ± 1 per cent is desired, the regulator illustrated in Fig. 8-28 for a series pass tube is adequate. However, if regulation of ± 0.1 per cent is required, a more complex regulator is required (5).

The circuit of Fig. 8-29 illustrates a shunt regulator which can be used for regulating the screen voltage to an amplifier tube. The regulation capability is of the order of ± 2 per cent.

8-10. D-C to D-C Power Supplies. Mobile transmitter applications require the use of a power supply capable of converting the low-voltage d-c battery supply to a higher d-c plate voltage for the transmitting

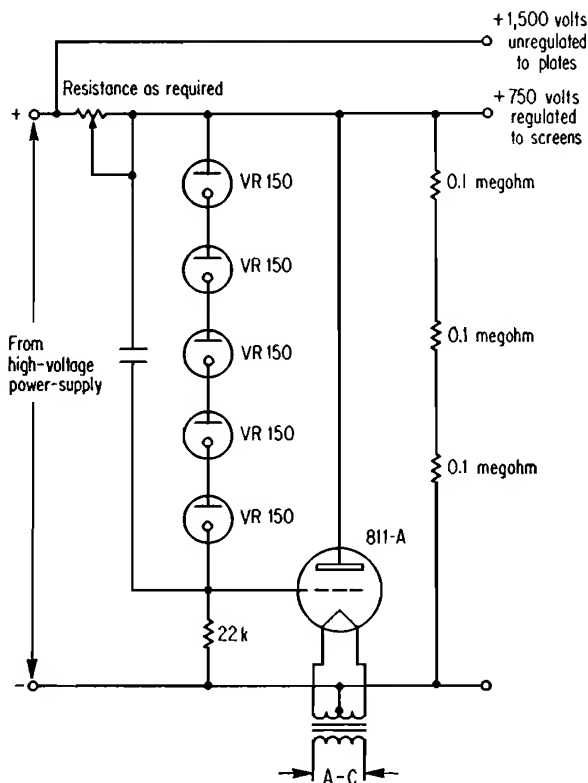


FIG. 8-29. Typical shunt regulator for supplying a screen grid.

tubes. This can be done mechanically with a dynamotor or a vibrator type of power supply; or it can be done electronically with a transistor converter.

The Dynamotor. The dynamotor is basically a motor and a generator wound on one armature, with both the motor and the generator using a common field winding. Because of the relatively high cost of the dynamotor compared to the vibrator type of power supply, the dynamotor is used mainly for the higher-power applications. The switching ability of the vibrator at high currents and high voltages is somewhat limited.

The dynamotor requires some occasional maintenance. There are four brushes in the machine: two for the driving motor and two for the generator. These will require occasional replacement. Similarly, the bearings must be occasionally lubricated. After prolonged use, the commutator will be worn because of brush friction. In such cases, the commutator will have to be turned down on a lathe. In spite of these disadvantages, the dynamotor is a rugged machine, and in many cases there is no better device for the purpose. The over-all conversion efficiency of the dynamotor is generally 50 to 60 per cent.

The output ripple frequency of the dynamotor is high, so that in most cases the filter required is nothing more than a 2- to 8- μ f capacitor across the d-c output. Radio-frequency interference is sometimes caused by the brushes sparking. This is eliminated by connecting a 0.001- μ f capacitor from the high-voltage brush to ground.

The Vibrator Power Supply. The vibrator, which is the heart of the vibrator supply, is basically an electrically driven vibrating reed. This reed has contacts which are arranged to switch the primary currents to the transformer, and in some cases, the secondary voltages too. The switching arrangement of the vibrator depends on the circuit. There are two general types of vibrator power supply, i.e., the nonsynchronous and the synchronous. The nonsynchronous type of vibrator supply requires the use of a rectifier to obtain a d-c output from the supply. Some typical nonsynchronous vibrator supplies are shown in Fig. 8-30. Each one of these supplies uses a different method of energizing the reed-driving coil. The advantage of the methods shown in Fig. 8-30a and b is that the coil current is not switched by the power contacts, and therefore good starting will be obtained at low battery voltages and at the end of the vibrator life. This is not true in the case of the simpler vibrator shown in Fig. 8-30c.

In all types of vibrator circuits, the power transformer primary is operated with a center tap as shown in Fig. 8-30. This prevents the core from being magnetized. The switching action of the vibrator continuously changes the direction of the current through the primary winding and hence the direction of the magnetic field in the core.

Some vibrator manufacturers use a shorted secondary over the vibrator energizing coil. This reduces the inductance in the energizing winding and hence reduces the sparking at the contacts. The life of a vibrator depends on the contact switching load, but it usually does not exceed 2,000 hr. The typical frequency of a vibrator supply is 100 to 120 cps.

The synchronous vibrator supply has an additional set of contacts on the reed which are used to switch the transformer secondary voltage in synchronism with that of the primary. In this way, the polarity of

the output is kept the same, and the need for a rectifier is eliminated. A typical synchronous vibrator supply is shown in Fig. 8-31.

This type of supply results in a more compact unit since the rectifier is eliminated. However, since another set of contacts is required the

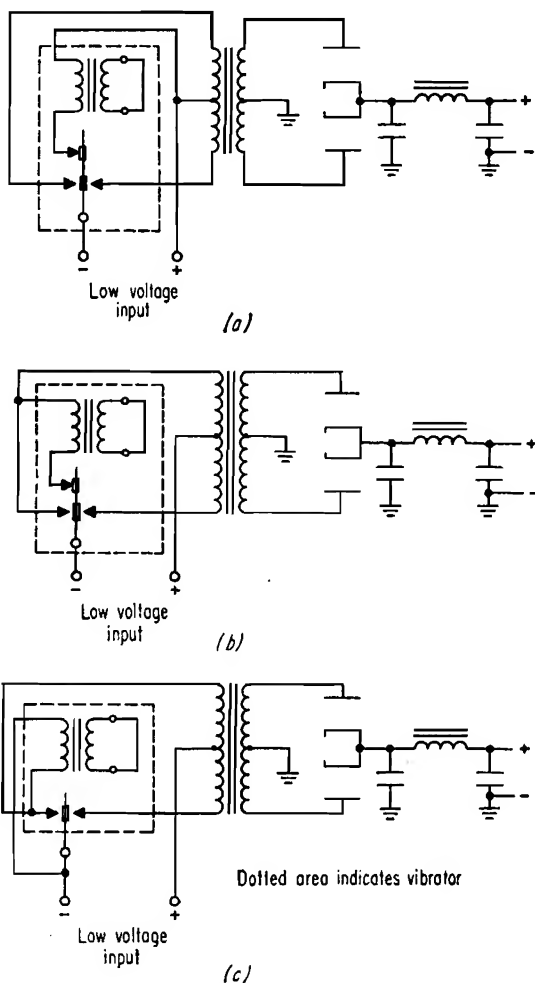


FIG. 8-30. Three types of nonsynchronous vibrator power supplies.

reliability of this type of vibrator is generally less than that of the non-synchronous types.

The Buffer Or Timing Capacitor. This is the capacitor across the power transformer secondary. The purpose of this capacitor is to make the voltage across the vibrator contacts zero or nearly so when the vibrator is switching. If the voltage across the contacts is zero during

the switching time, the contact life will be extended, since the only wear will be mechanical. Figure 8-32 gives the vibrator voltage waveform. The dotted curve gives the damped oscillation that results from the addition of the buffer capacitor to the power-supply circuit. This

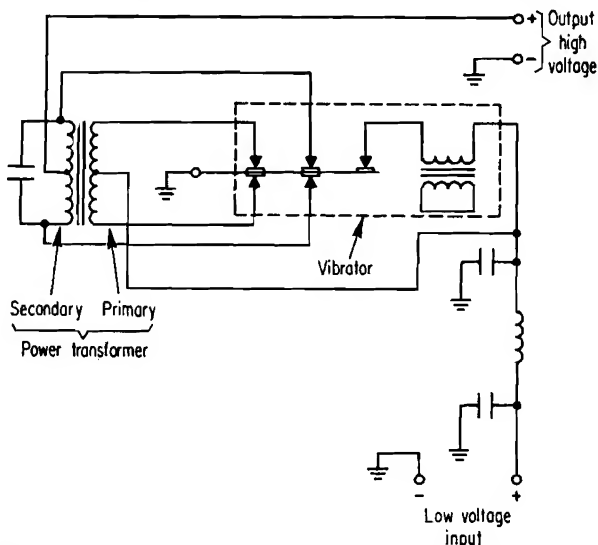


FIG. 8-31. Basic circuit of a synchronous or self-rectifying vibrator power supply.

oscillation occurs when the d-c current flowing in the primary is suddenly dropped to zero. The oscillation is actually suppressed at A when the vibrator contacts connect the battery to the other half of the transformer primary. The frequency of this oscillation is determined by the value of the buffer capacitor. The value should be such that the voltage across the contacts will be essentially zero when the contacts are made. This is the condition shown in Fig. 8-32. If the value of the buffer capacitor is more or less than the optimum value, the vibrator life will be shortened.

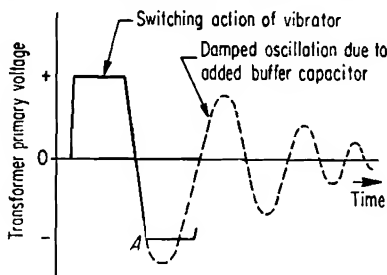


FIG. 8-32. The effect of a buffer capacitor on vibrator switching.

Transistorized Power Supplies. The typical transistor power supply uses two power transistors in a push-pull oscillator or multivibrator arrangement. This is shown in Fig. 8-33. The choice of the frequency of oscillation is limited by the upper-frequency limit of the transistor and by the transformer core losses. Transformer core losses increase as the square of the frequency. The optimum balance of core loss and

transformer size is in a broad region around 400 cps. For this reason, transistor-operated power supplies operate from 1,000 to slightly below 400 cps for the larger units. The higher oscillator frequencies result in a higher ripple frequency in the output, thus reducing the size and the complexity of the d-c filter. The actual operating frequency of the transistor-operated supply is determined by the primary turns, the transformer core loss, the cross-sectional area of the core, the flux density of the core, and the primary supply voltage.

Most transistor power supplies cannot be damaged by output short circuits or overloads. The effect of overloading a supply of this type is simply that the oscillator stops. Thus the unit is self-protecting.

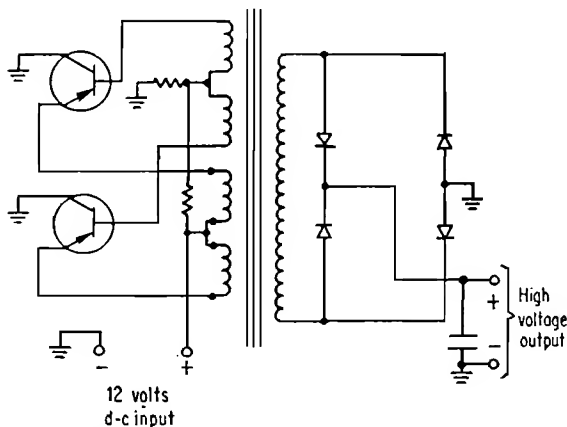


FIG. 8-33. A typical transistorized power supply.

The efficiency of this type of supply is approximately 70 to 80 per cent when supplying the designed load. Load currents more or less than the designed load result in lower efficiencies. Since there are no moving parts in a transistor-operated power supply, the life is limited only by the quality and rating of the components themselves. The life of this supply can easily extend for 50,000 hr with proper design and selection of components.

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CONTROL AND PROTECTIVE CIRCUITS

9-1. Control and Protective Circuits. The transmitter control and protective circuits act together to coordinate and protect the transmitter circuits and components. Basically, the control circuits are those which arrange for the systematic operation of the transmitter, while the primary function of the protective circuits is to prevent damage to the components and tubes in the transmitter. In practice, these circuits are very closely related and so are considered together in this chapter.

9-2. Functions of the Control Circuits. The functions of the control circuit can be subdivided into the following major types of circuits.

1. Power-control circuits for turning the transmitter power on and off
2. Transmitter-sequencing and interlocking circuits
3. Indication and alarm circuits
4. Transmitter-recycling circuits
5. Remote-control circuits
6. Automatic-tuning circuits

9-3. Power Control. The control of power circuits is accomplished by means of circuits utilizing *contactors*. The contactor is a magnetic switching device, identical in principle with the relay. However, the contacts and general construction are designed for switching power.

Contactors. Electrical contactor size has been standardized by the National Electrical Manufacturers Association (NEMA). The electrical characteristics for contactors from size 00 to size 7 are given in Table 9-1. Enclosed-contactor current ratings are lower since NEMA specifications limit the temperature rise of the contactor. The rating of a contactor is also reduced for lamp loads. The cold resistance of incandescent lamps is approximately one-tenth that of its hot resistance. This means that the contactor must initially carry up to 10 times its normal load when switching lamp loads. Thus the contactor rating is reduced to compensate for this high initial surge current. This rating applies only to those contactors switching tube filament loads.

Contactors are neither rated nor designed to interrupt or open short-circuit load currents. The normal contactor design permits opening

TABLE 9-1. CONTACTOR SIZES AND RATINGS

Contactor NEMA size	Current rating of contacts		
	Enclosed	Open	Incandescent lamp load
00	9	10	5
0	13	15	8
1	22	25	12
2	45	50	30
3	90	100	60
4	135	150	120
5	270	300	240
6	540	600	480
7	810	900	720

circuits with currents up to 10 times the rated current of the contactor. Circuit breakers or other protective devices should be used to open circuits with overload currents in excess of this amount.

The power poles of an NEMA-rated contactor are always *normally open*. For auxiliary control-circuit switching, the contactor may make use of interlock contacts.

These interlocking contacts, which are available in both normally open or normally closed designs, have current ratings of approximately 10 amp. These interlocking contacts are normally operated by the contactor mechanism.

Across-the-line Starting. The majority of transmitter power circuits are energized by connecting the load directly across the power line by the contactor. This load may be any of the transmitter power supplies or the motors used in the heat exchangers or blowers. Across-the-line starting is permissible whenever the starting current of the load is within the maximum short-time current ratings of the load itself and is capable of being supplied by the incoming power line. In addition, the across-the-line starting should not produce excessive voltage dip or transient voltages on the power line which may adversely affect the performance of other equipment operating off the same power line. If any of these effects occur, then some form of step starting must be used. A typical example of across-the-line starting is shown in Fig. 9-1.

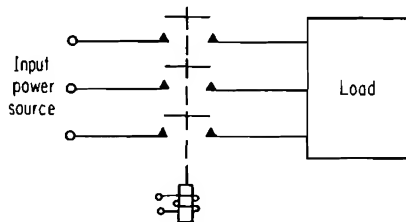


FIG. 9-1. Across-the-line starting.

Step Starters. A convenient method of reducing the incoming current surge when turning on very large power transformers or motors is to use a reduced-voltage starter or step starters. Basically, this method utilizes some form of current-limiting or voltage control, which is removed from the circuit in steps until the full power-line voltage is applied to the

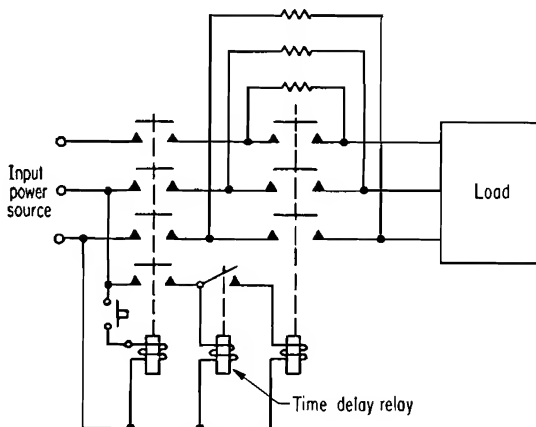


FIG. 9-2. A resistance step starter.

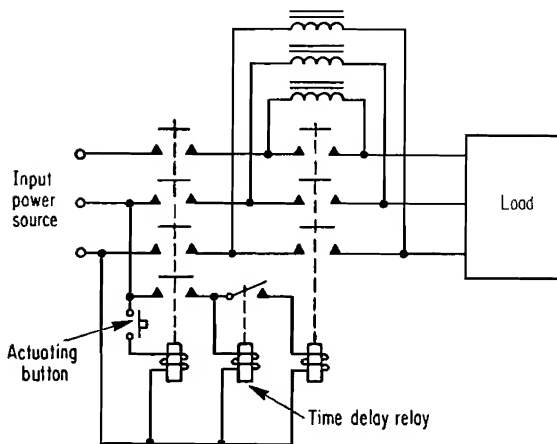


FIG. 9-3. A reactance step starter.

load. This voltage control may use series resistors or inductors or may make use of tapped transformers.

In the basic resistance step starter shown in Fig. 9-2, the resistance is shorted out after an interval determined by a time-delay relay.

An inductance can be substituted for the resistor as shown in Fig. 9-3 with the advantage that very little heat will be dissipated in the inductance. However, the size of the unit will increase considerably.

A third type of step starter is the autotransformer in which taps are changed to gradually increase the voltage applied to the load. (For the basic circuit see Fig. 9-4.)

The amount of current surge that can be tolerated will determine the number of steps used in the step starter. A step starter using one reduced-voltage point before the full-voltage tap is called a *two-point starter*. Similarly, a starter with two reduced-voltage taps is called a *three-point starter*.

9-4. Transmitter Sequencing and Interlocking Circuits. Except for the simpler low-power transmitters, the transmitter starting must be arranged to follow a specific order of events. In many transmitters, this starting procedure is made completely automatic without requiring any

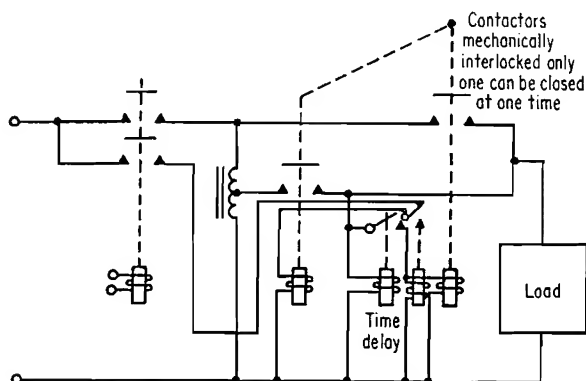


FIG. 9-4. An autotransformer step starter.

operations being performed by the operator. In other transmitters, the sequence of operations must be performed manually by the operator. Generally, the sequence of events cannot be interchanged owing to the interlocking of the control circuits. Interlocking is the common method used in transmitter control circuits to ensure that each step of the starting sequence is operating before the next step can be energized. Circuit interlocking is done by the interlock contacts on the power contactors for each of the transmitter power supplies. For example, when the filament contactor closes and energizes the transmitter filaments, the interlock contact on the filament contactor also closes. This is arranged to energize the time-delay relay, which, after the delay will allow the coil of the plate supply contactor to be energized. In this way, the proper sequence is secured. A basic example of interlocked circuits is shown in Fig. 9-5.

Starting Sequence. The basic starting sequence for transmitters using negative-grid grounded-cathode amplifiers is as follows:

1. The control circuit is energized.
2. The coolant blowers and pumps are turned on.
3. The coolant flow interlocks must be closed.
4. The transmitter filaments are turned on.
5. The high-voltage delay is energized.
6. The door interlocks must be closed.
7. The grid-bias power supplies are energized.
8. The plate supply is energized.
9. The screen-grid power supply is energized, if pentode or tetrode is used.

If the amplifier uses resistance grid bias, which depends on the r-f drive for proper biasing of the tube, then the r-f drive must be applied

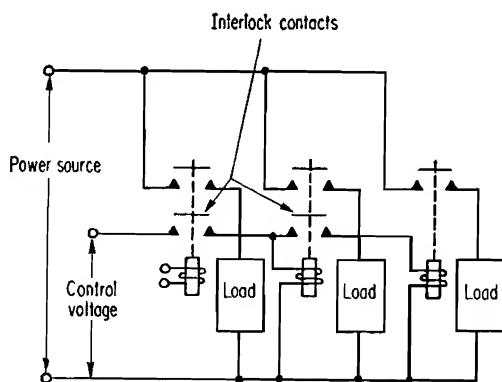


FIG. 9-5. A basic interlocked relay circuit.

as part of step 7 above. If fixed grid bias is used, then the r-f drive can be applied after step 7, 8, or 9 listed above.

For negative-grid tubes connected as an a-c grounded-grid amplifier, the r-f drive must be applied only after all the d-c voltages are applied to the tube. This protects the grid from damage resulting from the relatively large amount of r-f drive required for grounded-grid operation.

In transmitters using a klystron or traveling-wave tube as a power amplifier, the starting sequence is as follows:

1. The control circuit is energized.
2. The focus magnet power supplies are energized.
3. The coolant blowers and pumps are turned on.
4. The coolant-flow interlocks must close.
5. The klystron filament is turned on.
6. The beam high-voltage delay is energized.
7. The door interlocks must be closed.
8. The beam high-voltage power supply is energized.

Since the r-f drive has no effect on the current requirements of these tube types, it can be applied at any time during the sequence.

9-5. Indication and Alarms. The operating status of the control circuit of the transmitter is usually denoted by indicator lamps and alarm buzzers. Indicator lamps monitor the various operating circuits and will assist in the maintenance of the equipment in time of failure. The alarm buzzers are usually used to denote any transmitter failure, or any potentially dangerous condition.

Indicator Circuits. Indicator lamps may be located in the power circuits of the transmitter to indicate whether the circuit is energized. This method, while certainly the simplest and most foolproof, has a number of limitations which generally make it a poor choice in all but the lowest power applications. Locating the indicator lamps in the power circuits requires that the wiring to the indicator lamps be capable of carrying the current rating of the circuit in which the lamp is located without burning up. In this way, the wiring is protected in case of a fault in the indicator lamp wiring. However, if the lamp is in a 100-amp circuit, it becomes impractical to have 100-amp wiring just for the indicator lamps. A second fault of this method is that, if the circuit across which the indicator lamp is located is inductive, the collapse of the magnetic field when the circuit is opened may cause an excessively high voltage to appear across the lamp. This may not only burn the lamp out, but may also cause the lamp holder to break down.

A second method, and a superior one, is to have a separate circuit for the indicator lamps operated by a separate interlock contact on the contactors and control-circuit relays. This avoids the two difficulties described above. A basic example of this system is shown in Fig. 9-6.

Indicator Lamps. Indicator lamps may be of the neon or the incandescent type. The neon lamps have the advantage of long life and a mechanical ruggedness due to the absence of a filament. However, the intensity and the total light output of the neon is low. For example, the candle power of a $\frac{1}{4}$ -watt neon (type NE48), which is the largest type generally used for indicator service in transmitting equipment, has a total light output of 0.01 cp (candle power). This is 40 times less than the light output from the incandescent type 47 indicator lamp, which requires less than 1 watt input.

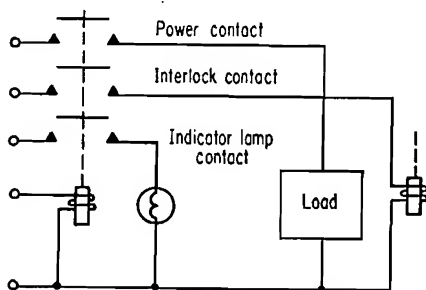


FIG. 9-6. Separate contacts for the power, interlocking circuit, and indicator lamp circuit avoid many problems discussed in Sec. 9-5.

The life of a neon lamp will vary from 7,500 to over 25,000 hr. This life depends on the type of lamp and on the value of current-limiting resistor in series with the lamp. In the screw-base neon lamps there is no choice, for the resistor is built into the base of the lamp. However, in the bayonet-base neon lamps, this current-limiting resistor must be added by the user. If the value of the resistor is lowered to obtain a greater light output, then the life of the lamp is shortened. This is shown in the typical curve of Fig. 9-7.

Neon lamps do not burn out. Rather, the bulb gradually becomes darker and blackens until the light output is so low as to be unusable. The light from a neon lamp consists of a reddish orange glow. For this reason, the neon lamp cannot be satisfactorily used in indicators with green or blue jewels.

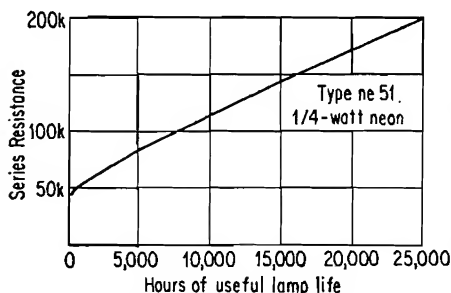


FIG. 9-7. The effect of the series current-limiting resistor on the life of a neon lamp.

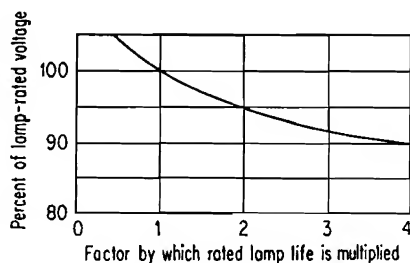


FIG. 9-8. The effect of lamp voltage on life for incandescent lamps.

The life of an incandescent lamp is very dependent on the ratio of the applied voltage to its designed or normal voltage. The higher the filament voltage, the higher will be the filament temperature, and the faster will be the rate of the evaporation of the tungsten filament. The curve of Fig. 9-8 gives the approximate relationship between the applied filament voltage and the life expectancy of an incandescent lamp. Thus theoretically, by undervolting a 3,000-hr lamp by only 10 per cent, the life of the lamp can be extended to 12,000 hr. However, in practice, the filament of a lamp crystallizes with use. This makes the filament very brittle and subject to breakage due to any small vibration or shock. For this reason lamp manufacturers guarantee incandescent lamps for a maximum of 5,000 hr. Lamps designed for this life span are often classed as having an "indefinite life." Once again, this does not mean that the lamp will last indefinitely, but rather that the lamp represents the best possible combination of all the factors, both electrical and mechanical, that make for long lamp life.

Alarm Circuits. Alarms are used to audibly indicate a transmitter failure and as a warning for a potentially dangerous condition. This

latter condition includes low liquid-coolant level, high air temperature, high coolant temperature, or low r-f output or input. In a properly interlocked transmitter, any serious failure that effects the transmitter performance will inevitably cause the plate contactor to open. Thus, a normally closed interlock contact on the plate contactor can be utilized to energize an alarm. An example of this technique is shown in Fig. 9-9.

To prevent the continuous sounding of the alarm in the event of a transmitter failure, the alarm usually has some means of disabling the alarm buzzer. This can be accomplished simply by placing a switch in series with the alarm buzzer. However, this also requires that the operator remember to place the alarm switch back on when the equipment is again operating. For this reason, many transmitters make use

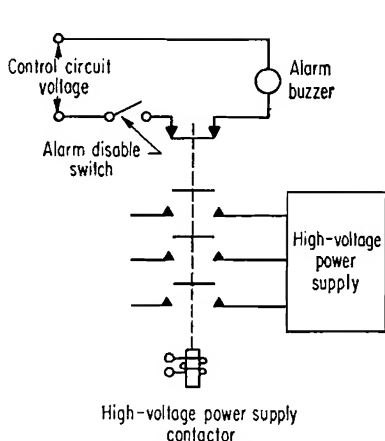


FIG. 9-9. Basic alarm circuit.

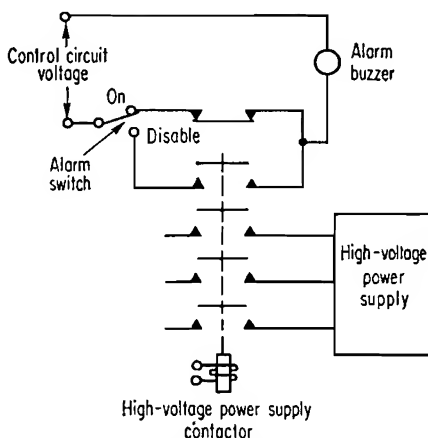


FIG. 9-10. Self-reminding alarm circuit.

of a "self-reminding" circuit similar to that shown in Fig. 9-10. In this circuit, the alarm may be disabled after a fault. However, after the fault has been cleared, the alarm will again sound. This will remind the operator of the transmitter to place the alarm switch back into the alarm-ready position.

9-6. Transmitter Recycling Circuits. Many causes of transmitter failure are only temporary conditions which are cleared almost immediately after the failure. Examples of these faults are flash-arcs in tubes, arc-over on dirty insulators, and lightning striking the antenna. To minimize lost transmitter operating time, many transmitters utilize automatic-recycling circuits to restore the transmitter to operating condition as soon as possible after an overload. These recycling circuits are usually arranged to allow the transmitter to be reenergized from one to three times. If the fault is not a temporary one of the type described, the transmitter will try again and fail until the number of recycling

attempts are used up. The transmitter will remain off until the fault condition is corrected.

A basic recycling circuit is shown in Fig. 9-11. This circuit is based on a mechanical *ratchet* or stepping relay. The operation of this circuit is as follows:

1. A d-c overload through the coil of relay *A* causes the normally closed contact of relay *A* to open, thus deenergizing the coil of contactor *F*.

2. A d-c overload also causes the normally open contact of relay *A* to close and energize the coils of relays *B*, *C*, and *E*.

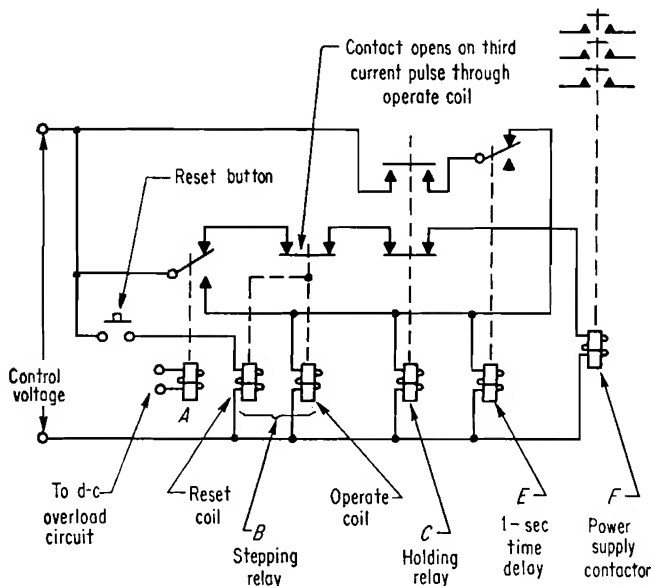


FIG. 9-11. A basic recycling circuit.

3. The normally closed contact of the time-delay relay *E* keeps the coils of relays *B*, *C*, and *E* energized for a period of the time delay of relay *E*.

4. The contacts of relay *C* are open, which ensures that the power contactor does not close for the period of the time delay.

5. At the end of the time-delay period, the power contactor *F* again closes. If an overload again is present, the relay *A* is again energized, and the cycle is again repeated.

6. If a third overload occurs, the normally closed contact of the stepping relay *B* opens. This prevents further energization of the contactor *F* until the stepping relay is reset by pressing the button *D* which energizes the reset coil of the stepping relay *B*.

A second basic type of recycling circuit is shown in Fig. 9-12. This

circuit is based on the use of individual relays in a progressive lock-up circuit. This circuit will permit three recycles before the transmitter is disabled. The operation of this circuit is as follows:

1. Normal operation of the high-voltage on switch will operate the power contactor *A* through the normally closed contacts of the d-c overload relay and through the normally closed contacts on relay *B*.

2. The normally open interlock contact on the power contactor *A* will close and lock the relay *G* closed through the normally closed contacts on relays *D* and *F*.

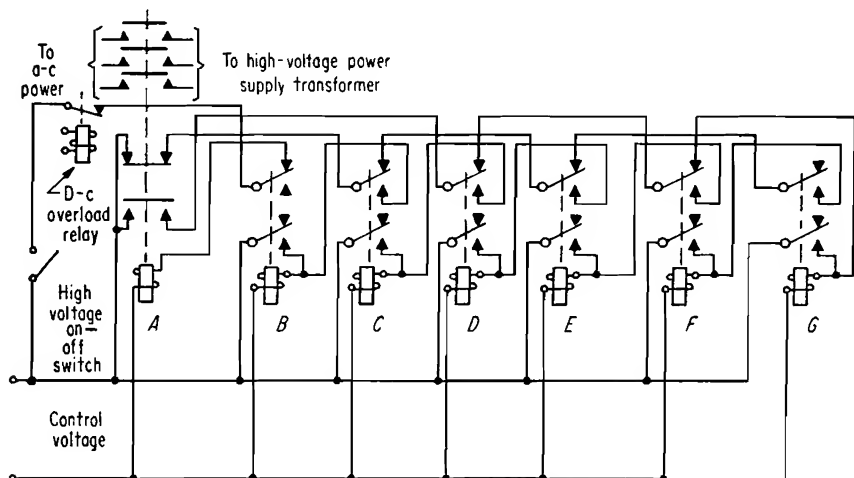


FIG. 9-12. A recycling circuit using a progressive relay lockup circuit.

3. If a d-c overload occurs, the normally closed contact on the d-c overload relay will open and cause the power contactor *A* to open and also cause the normally closed contact on *A* to again close.

4. This will result in the relay *F* becoming energized through the normally closed contacts on relays *C* and *E* and the locked relay *G*. This will lock relay *F* closed through its own contacts.

5. Each time a reclosing and reopening of the power contactor *A* occurs, another relay will lock closed in the manner described.

6. On the third opening of the power contactor *A*, the relay *B* will close, thus opening the circuit to the coil of the power contactor *A*, thus preventing further reclosures of the contactor until the entire circuit is unlocked. This is accomplished by opening the power source to the relays.

7. Normal opening and closing of the high-voltage on switch does not cause the relays to lock. The second pole of the double-pole high-voltage on switch opens the circuits to all the relay coils. This effectively resets the circuit each time the switch is turned off.

9-7. Remote Control. Remote control of commercial transmitters is under the jurisdiction and control of the FCC. The method of achieving this control is not specified by the FCC as long as it will provide for the following control and metering functions:

1. Turn the filaments off and on.
2. Turn the transmitter off if the control system fails in any way.
3. Turn the transmitter final plate supply off and on.
4. Meter the final output tube plate current and plate voltage.
5. Meter the antenna current.
6. Monitor the frequency and modulation level.
7. Monitor the proper operation of the tower beacon and obstruction lights.

The FCC requirements ensure that the transmitter is operating within the legally assigned power limit and is under the operator's control.

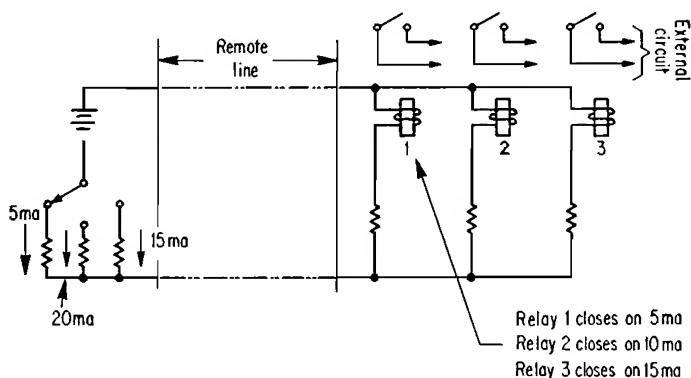


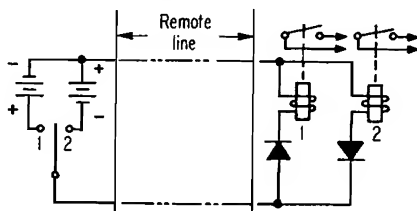
FIG. 9-13. Current differential method of remote control.

This is the extent of the interest the FCC has in the operation of the transmitter; it does not cover the multitude of other functions that are necessary to ensure a long, unattended life for the transmitter. Some of the other desirable control functions are to meter exciter currents, to meter final amplifier screen and control-grid currents, to meter the final filament voltages, to tune various r-f circuits, to raise and lower filament and plate voltages, to act as fire alarm, to reset after unauthorized entrance and overload.

Many methods of remote-controlling transmitters are available. However, the methods are so numerous that only the basic systems can be discussed. Remote-control systems can be roughly divided into d-c and a-c systems. D-c control systems are generally simpler than a-c systems; however, d-c systems are generally not compatible with the systems of most commercial telephone companies. D-c control signals are also not capable of being transmitted through microwave links.

D-C Remote-control Systems. One basic d-c system is called the *current-differential system*. Each signal in this system corresponds to a different current level. Each of the current levels will operate a relay at the receiving end in the manner shown in Fig. 9-13.

For example, the first relay may close at 5 ma, the second relay will close at 10 ma, the next at 15 ma, etc. This method is satisfactory for starting sequences where each relay must operate in turn.

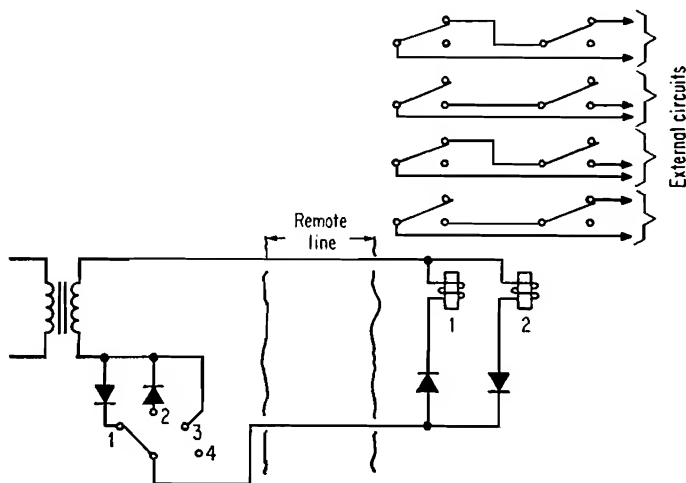


When switch is in position 1, only relay 1 will close.

When switch is in position 2, only relay 2 will close.

FIG. 9-14. Polarity-sensitive relay method of remote control.

Another basic d-c system is based on voltage polarity. A change in polarity across a pair of wires will select one of two relays to operate as shown in Fig. 9-14. These relays can either be standard d-c relays with rectifiers polarized in the manner shown in Fig. 9-14, or they can be



When switch is in position 1, relay 1 closes.

When switch is in position 2, relay 2 closes.

When switch is in position 3, both relays close.

When switch is in position 4, both relays open.

FIG. 9-15. Polarity-controlled remote control where either of four conditions may be selected.

magnetically polarized relays which do not require the series rectifiers. If the battery sources are replaced by half-wave rectifiers, as shown in Fig. 9-15, then either relay may be closed independently, as well as both simultaneously. If one includes the condition when both relays are

open as constituting a control operation, then four separate functions can be achieved over a single pair of wires. Two of these systems can be used with ground as a common return, in which case 16 separate functions

can be executed over a single pair and ground.

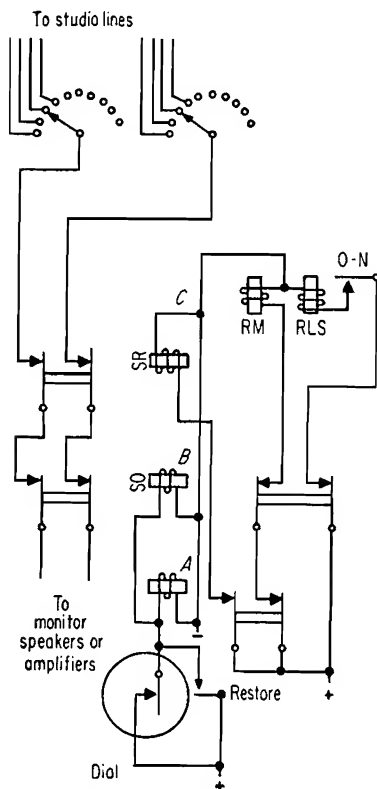


FIG. 9-16. Remote-control system using d-c pulses. Contacts closed when rotary magnet off normal position; O-N, off-normal springs. (Courtesy of Radio-News.)

use tones, either in a pulsed or continuous form, to couple the controlled function at the transmitter site to the controlling operation at the remote location. At the transmitter end, the audio tone can be rectified and used to operate a stepping relay in the same manner as that described for the d-c control systems.

To permit more versatility in the a-c system, many audio tones are often coupled to the same pair of lines. Selective filters are used at the transmitter end to differentiate between the channels. Two systems of this type are shown in Figs. 9-17 and 9-18. The system shown in Fig. 9-17 uses a highly selective filter made up of inductances and capaci-

The third basic d-c system is to use d-c pulses generated by a dial switch to actuate a stepping relay as shown in Fig. 9-16. The operation of this circuit is as follows: each time the dial switch closes, the relay *A* operates. This relay in turn operates the rotary magnet *RM*, to step the wipers, and relay *C* to open the wiper circuit during rotation. Relay *C*, being slow to release, remains operated until the end of the series of dial impulses. Relay *B* is sufficiently slow to operate to remain normal during the transmission of the short dial impulses. However, the longer manual closure of the restore key operates relay *B* to actuate release magnet *RLS* and release the switch. Break contacts on relay *B* again open the wiper circuit during rotation to guard against the previously mentioned breaking of currents by the wipers as well as to avoid the momentary energizing of the nonselected circuits.

A-C Remote-control Systems. A-c remote-control systems are an extension of the simpler d-c systems previously described. The a-c systems generally

tances. This requires the use of amplifiers following the filter to make up for the insertion loss of the filter. The method shown in Fig. 9-18 uses a resonant reed relay as the selective device. The reed which also forms the armature of a relay will vibrate only when excited by its

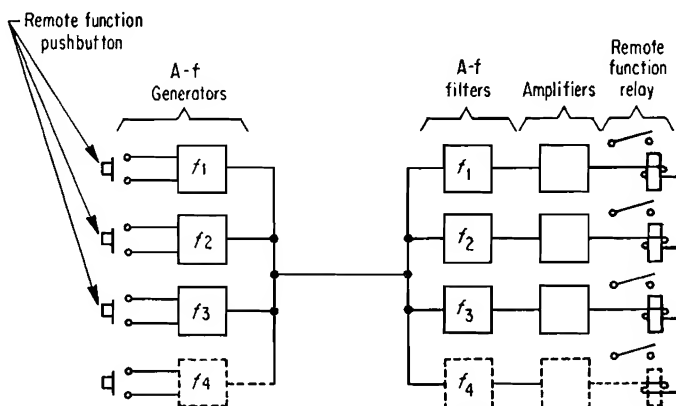


FIG. 9-17. A-c remote control based on audio tones.

resonant frequency. The rapidly vibrating contacts of the reed relay are used to operate an auxiliary relay which will do the actual circuit controlling or switching.

9-8. Transmitter Telemetering. The FCC requires that the antenna r-f current, the final plate voltage, and the final plate current of the

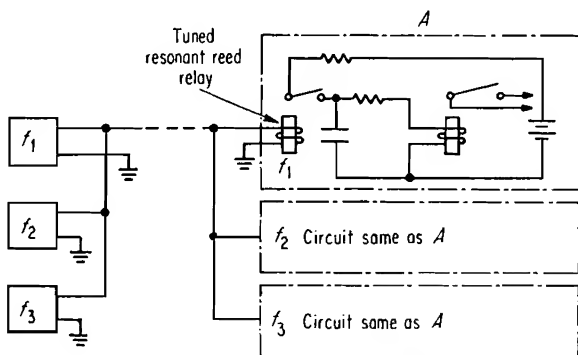


FIG. 9-18. A-c remote control based on resonant reed relays.

remote transmitters be measured at the operating location. In d-c remote-control systems, the d-c sampling method shown in Fig. 9-19 may be used. This system is calibrated by first selecting a known voltage at the remote location with the stepping switch. This known voltage can then be used to calibrate the meter at the operating position. This

prevents errors in meter readings due to variations in the line resistance between the transmitter and the operating location.

A-c remote-control systems can make use of the system shown in Fig. 9-20. A d-c sample of the voltage or current to be metered is

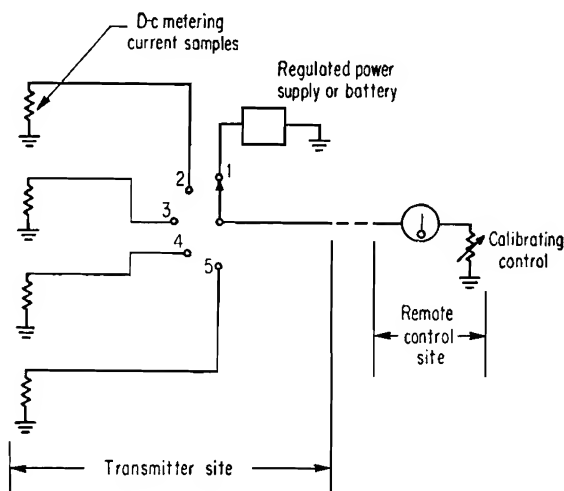


FIG. 9-19. Basic d-c method of telemetering transmitter operating conditions.

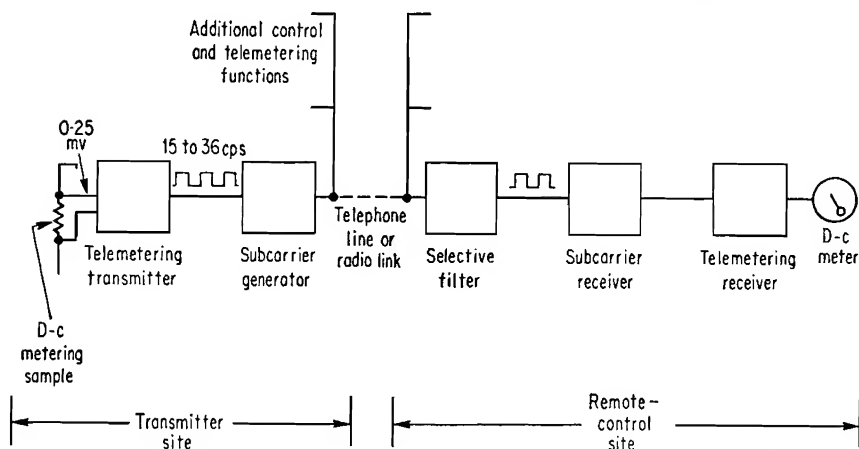


FIG. 9-20. Basic a-c method of telemetering transmitter operating conditions.

applied to a telemetering transmitter. This is a device that takes the voltage or current sample and produces a frequency output according to the level of the input d-c signal. The typical input to the telemetering transmitter is 0 to 25 mv, and the output frequency is usually a train of square waves between 15 and 35 cps.

While this audio frequency contains all the metering information, the

low frequency of the square waves does not lend itself to satisfactory transmission through most systems and telephone lines. Therefore, the square waves are usually used to key a high-frequency tone off and on. This higher-frequency tone is known as the *telemetering subcarrier* and is usually in the 2- to 3-ke region. At the operating location, the telemetering subcarrier is applied to the telemetering subcarrier receiver. The 15- to 35-cps output signal is then applied to the input of a telemetering receiver. This is basically a frequency counter. This receiver will cause a meter to deflect according to the frequency of the input.

There is an almost endless number of transmitter telemetering systems, each of which is tailored to some individual application. However, they all perform the same general function as the system previously described. For example, a d-c current-metering sample can be used to control the frequency of an oscillator operating in the vicinity of 30 ke. This can be used to modulate a transmitter in f-m broadcast service. This 30-ke subcarrier can be detected by a special receiver at the remote-control site. The amount of deviation of the subcarrier oscillator is proportional to the sampled meter reading at the transmitter. Thus this system makes use of the transmitter itself to serve as the telemetering link and requires no additional telephone lines for this purpose.

9-9. Automatic Tuning of Transmitters. Automatic tuning is a feature usually found only on transmitters either intended for operation by inexperienced personnel or where an immediate selection of a large number of operating channels is required. Mobile marine and aircraft transmitters are typical applications.

Automatic tuning can be subdivided into two general methods, i.e., those systems in which the transmitter can be preset to operate in any of a number of preset channels, and those in which the transmitter tuned circuits are controlled and adjusted by a servomechanism sensing some parameter related to the adjustment of the tuned circuit. The sensing may be based on the transmitter frequency, the amplitude of the voltage across the tuned circuit, or the r-f match produced by the tuned circuit. Automatic systems of this type are normally quite complex and are not commonly used unless the intended service of the equipment thoroughly justifies this means.

In one of its simplest forms, the present method of channel selection may make use of a selector switch to directly select a preset inductance or capacitance. Such a selector switch may be manually set by the operator, or it may be electrically positioned by a device called a *discrete positioning actuator*. A more complex type utilizes a mechanical positioning head called an *Autotune*, made by the Collins Radio Company. In this device, the adjusted component is mechanically positioned to a preset position which has been previously selected by the operator.

9-10. Protective-circuit Functions. The purpose of the protective circuit is, as its name implies, to prevent extensive damage from an overload or malfunction of a component within the transmitter. This overload can be the result of a defective tube or other component within the transmitter, or it can be the result of an external fault such as an antenna failure, lightning, or r-f transmission-line arc-over. Thus the protective circuits are often required to sense a-c, d-c, and r-f overloads. The protective circuits are also interlocked with the control circuits to prevent damage to other components when one circuit is deenergized by the protective circuit. An example is this: If an overload occurs in the plate circuit of a tetrode, the screen-grid supply must also be deenergized to protect the screen grid from damage due to the absence of plate voltage.

9-11. Protective-circuit Philosophy. The principle purpose of any protective circuit is to prevent fires. Regardless of the type or extent of the protective circuitry in a piece of equipment, an overload should not allow any component or wire to reach a temperature that will result in a fire.

This rule still allows a considerable variation in the methods of achieving this protection. Much depends on whether it is desired to protect the wiring or the component. If only the wiring is protected, the component must be considered expendable. This is often the case with the smaller tubes. In general, the amount of protection given to any component depends largely on its cost. One will rarely find a 25-dollar circuit breaker protecting a 2-dollar tube. If the extent of any damage that will result from the failure of this tube is held only to the tube, then the cost of a circuit breaker can hardly be justified under normal circumstances. For this reason, the exciter tubes and other low-level tubes in a transmitter generally have no protection other than the a-c overload circuit breaker in the primary of the power supply feeding these tubes. In cases where the final tube is worth hundreds or even thousands of dollars, the complete protection of the tube against any overload, which might result in damage to the tube, is completely justified economically. Tubes of this type usually have an overload device in series with every element in the tube.

In observing the "*no fires*" rule, consideration must be given to the maximum overload current that can flow under short-circuit conditions, and to whether the fuse or circuit breaker can safely open this current. This rating is many times the normal current rating of the protective device and is generally available from the manufacturer's literature for the device. Under short-circuit conditions, the current is limited by the internal impedance of the generator supplying the load, the impedance of the power line, the regulation of any intervening power or distribution transformers, and the resistance of the short circuit itself. If the breaker

or fuse is not capable of opening this current, the current may continue to arc across the contacts after the breaker or fuse opens, or in the case of a circuit breaker it simply may not open at all. In either case a fire is certain to follow. For this reason, it is necessary to back up or precede the breaker or fuse that has an inadequate short-circuit rating with one that has an adequate short-circuit current rating.

Protective-circuit Speed. The speed of the overload protective circuit is of considerable importance in many transmitter circuits. In circuits using components that are marginally rated and where the components cannot sustain prolonged overloads, the speed of the protective circuits must be fast enough to prevent permanent damage. Marginally rated components are not characteristic of poor design, for, in many of the larger and high-power transmitter circuits, the choice of components is very limited. This is particularly true of power tubes. Furthermore, in higher-power applications the cost often increases sharply as the rating of the component increases. Thus generous derating of components is an extremely expensive method of achieving component protection. It is more economical to provide a fast-acting protective system for the components than to derate the components to take the expected overload currents.

To prevent permanent damage to a component where this condition potentially exists, it is necessary to (1) detect the overload current before it rises to the danger point, and (2) deenergize the source of power to the overloaded circuit before it can do permanent damage.

The required speed of this protection will vary according to the application. Many components, such as motors and transformers, have an initial starting current surge. This current surge is normal and is permitted by the design of the components. Protection for these applications requires an inverse time relationship between the time the circuit is deenergized by the protective circuit and the magnitude of the overload current. Thus the bigger the overload, the faster the circuit will be deenergized. This characteristic of the protective device permits the starting surges; but in case of severe overload currents the circuit will be disconnected from the power line very rapidly.

The list of protective devices typically used in transmitting equipment includes fuses and circuit breakers. More elegant circuits use current-sensing relays which are arranged to trip remotely located circuit breakers in the a-c line or vacuum switches in the d-c circuits. In critical circuits where the discharge of the d-c power-supply filter capacitor through the fault can cause damage, an electronic crowbar circuit can be used. This device electronically short-circuits the d-c supply voltage. All these components and circuits will be discussed in detail in the following sections.

9-12. The Fuse. General. The simplest method of both detecting an overload and disconnecting the circuit is by means of a fuse. Fuses come in a variety of characteristics to fit the requirements of almost any circuit. Most fuses fit into three general classifications, i.e., the fast fuse, the medium-lag fuse, and the delayed-action fuse. The time-overload curves for these three types of fuse are shown in Fig. 9-21.

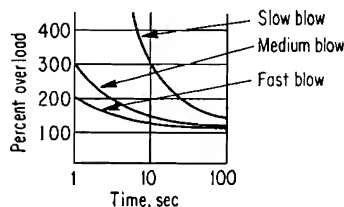


FIG. 9-21. Typical fuse characteristics.

Fast Fuses. This type of fuse is primarily used for the protection of meters as well as germanium and silicon rectifiers. They are available in ratings as low as 0.002 amp. The fast action of these fuses is given by the small thermal mass of the fuse element.

Medium-lag Fuses. This is the most commonly used fuse, comprising about 80 per cent of all the fuses used by the electronic industry. It is used for transformer primary and secondary protection and for plate-circuit protection of tubes. A zinc wire or strip is usually used for the elements in this type of fuse.

Delayed-action Fuses. The delayed-action fuse is used in circuits having a high starting current—such as is encountered in motor starting, in incandescent-lamp loads, and for starting tube filaments. This type of fuse is often favored where the normal current is close to the blowing point of the fuse and where continuous load changes result in cyclic failure of the simpler types of fuses. Cyclic failure of a fuse is due to the continuous mechanical stresses set up by the expansion and contraction of the fuse element with current changes. This eventually results in failure of the fuse even though the fuse was not subjected to current overload.

Fuse Ratings. A fuse is generally capable of carrying 110 to 115 per cent more than its current rating indefinitely. This same fuse is normally designed to blow in 1 hr when carrying 135 per cent of the normal rated current. However, a fuse that is operated at 100 per cent of its normal rated load will generally fail in 1,000 to 3,000 hr. The safest method of rating a fuse to obtain the longest fuse life is to use the highest current rating that can be tolerated by the circuit and its components and still obtain the desired protection. For short-circuit protection, fuses loaded to 25 to 50 per cent of their rating will give a safe margin against cyclic fatigue of the fuse as well as brief surge currents.

Since a fuse is a thermally operated device, the current-carrying capacity of the fuse is modified by the ambient temperature of the air. Fuse ratings are generally based on an ambient temperature of 70 to 80°F. At temperatures above this, the effective current rating of the fuse is reduced. Conversely, the colder the ambient temperature, the higher

the effective current rating of the fuse. This is often an advantage in some circuits. The current ratings of motors and transformers are limited by the maximum temperature of the unit. With colder ambient temperatures, these components can have a greater temperature rise. Thus the protection offered by the fuse can be matched to the operating characteristics of these components.

9-13. The Circuit Breaker. There are two general types of circuit-breaker-actuating mechanisms, i.e., the thermal and the magnetic. The thermal circuit breaker depends on the generation of heat to operate a bimetallic strip which will trip the circuit-breaker mechanism. For this reason the thermal circuit breaker is a relatively slow-acting device.

The magnetic type of circuit-breaker mechanism, on the other hand, is capable of rapid response to overloads. A magnetic circuit breaker

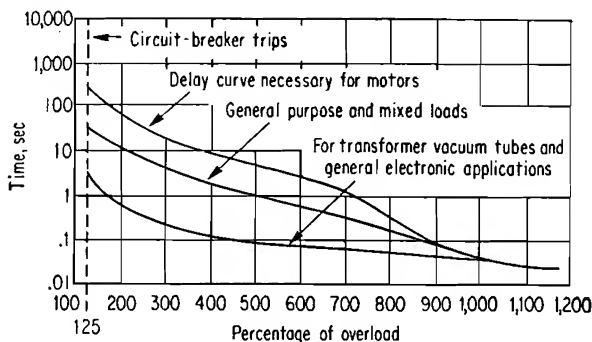


FIG. 9-22. Typical delay curves for magnetic circuit breakers.

can trip or open an overloaded circuit in as little as 10 msec. The characteristics are often modified to provide slower tripping time to allow for motor starting surges and transformer inrush currents. The typical response times for some magnetic circuit breakers are shown in Fig. 9-22.

Most of the molded-case circuit breakers combine both the magnetic and thermal trip mechanisms in one breaker. This type of circuit breaker is rated according to current capacity by letters starting with *F* and progressing through *L* for the largest size. The thermal trip device offers protection against small but long-term overloads, while the magnetic trip offers fast breaker action for high-current overloads. The magnetic trip mechanism is usually set to trip at 5 to 15 times the continuous current rating for this type of breaker. Up to this current, the protection is through thermal means.

9-14. Circuit-breaker Applications. Much of the versatility of the circuit breaker can be realized by separating the contacts and the coils in a circuit. In this way, the coil can be used to sense a d-c overload,

while the contacts can be placed in the a-c circuit of the power supply feeding the overload current. Many of the circuit combinations for circuit breakers are shown in Fig. 9-23. In many applications, auxiliary

switches are also mounted on the circuit breaker and are used to turn off other circuits in the transmitter.

In large circuit breakers, the power required to trip the circuit breaker in this manner becomes rather large and would consume a large portion of the power in the operating circuit. In these cases, the circuit breaker is conveniently tripped by an auxiliary trip device, which is known as either an *undervoltage trip device* or a *shunt trip device*. The undervoltage trip device must be kept energized to prevent the circuit breaker from tripping. If the undervoltage trip device becomes deenergized, even momentarily, the breaker will trip. By connecting the contacts of a relay that is sensing the d-c circuits, as shown in Fig. 9-24, any overload that causes the relay contacts to open will also open the undervoltage trip device in the circuit breaker and cause it to trip. In this way a very small but critical current can be made to trip a circuit breaker controlling a large power supply that might cause greater damage if the overload condition was allowed to progress.

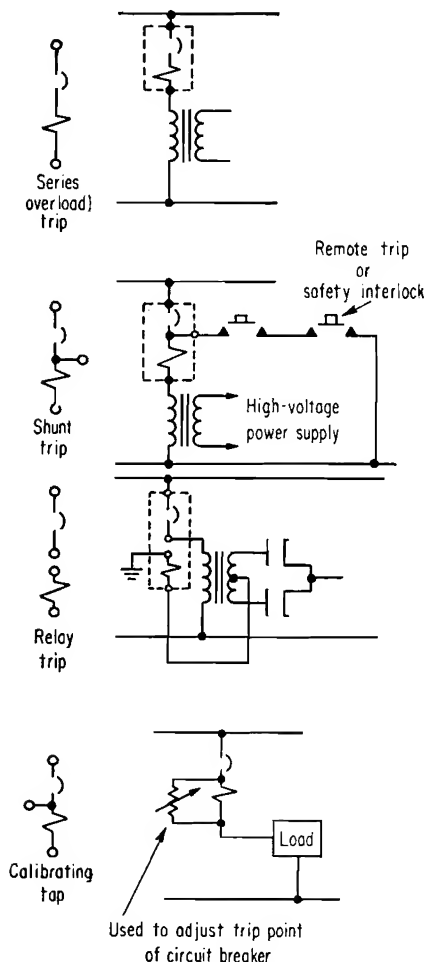


FIG. 9-23. Typical circuit breaker types and applications.

A circuit breaker can also use an auxiliary tripping system similar to the above except with a shunt trip device. This requires that a voltage be applied to the shunt trip device to trip the breaker. The disadvantage of this method is that it does not have the fail-safe feature offered by the undervoltage trip method just described. If the source of the shunt-trip-device energizing voltage fails due to the fault current, then the circuit breaker may never trip.

9-15. Electronic Crowbar. The short-circuit currents in high-power installations have been measured as high as 2,000 amp. Currents of this magnitude must be removed very rapidly to prevent extensive and permanent damage to the tubes and components in the transmitter. The electronic crowbar circuit described will successfully remove and deenergize the power supply in as little as $1\ \mu\text{sec}$. This is better than 8,000 times faster than the mechanical circuit breakers ordinarily used

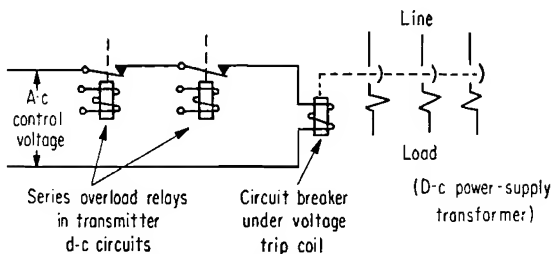


FIG. 9-24. Remote tripping of circuit breaker by means of an undervoltage release solenoid.

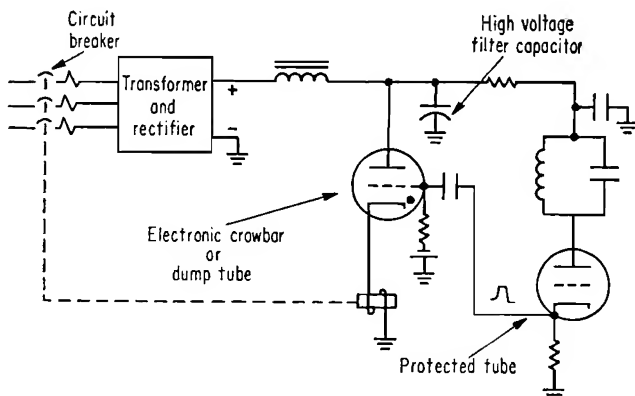


FIG. 9-25. Basic electronic crowbar fault protection circuit. (Courtesy of Electronics.)

for this application. The speed and effectiveness of the protection offered by the electronic crowbar are best illustrated by the fact that a piece of tin foil dropped across the terminals of a 7-kv power supply will leave only a small burn mark on the foil. This same experiment performed with the more conventional circuit-protection methods will result in the foil being almost completely consumed and vaporized.

The electronic crowbar circuit is shown in its basic form in Fig. 9-25. A thyatron is placed across the high-voltage power supply. Normally, this tube is not conducting. However, if an arback or other overload occurs in the tube, the voltage developed across a sampling resistor will

cause the thyatron to "fire." Gas thyatrons, when fired, present a very low resistance path. This effectively places a short circuit across the high-voltage power-supply terminals. This very rapidly discharges the stored energy in the filter capacitor. At the same time, the circuit breaker for the power supply is tripped, deenergizing the power supply.

In practical circuits, the method of detecting the fault current will vary. Some methods sample the current flowing through the high-voltage conductor before the thyatron protecting tube, thus providing protection for both the amplifier tube and the thyatron. To isolate the high voltage from the thyatron grid, an isolating transformer is often used. A fault or short circuit will result in a rapid change of current. This current change is passed through the transformer as a pulse which can be used to fire the thyatron. The design of this transformer is important if rapid response to an overload is to be expected. This rapid current change represents a high frequency which must be passed by the transformer.

The differential method of detecting opposite-going changes that occur in the plate current and the r-f output can be used to fire the thyatron. During a fault, the plate current will rise while the r-f output will drop. These two effects are applied to a differential amplifier, and the resultant signal is applied to the grid of the thyatron.

Some crowbar circuits have made use of grid-controlled rectifiers in addition to the shunt thyatron. When the thyatron fires, it is also arranged to block the rectifiers. With the rectifiers no longer conducting, the load current opened by the circuit breaker is very low. This prolongs the life of the circuit breaker.

9-16. R-F Overload Protection. Failure of the r-f components and circuits will often result in the progressive failure of the transmitting tube supplying the overloaded r-f circuit. Unfortunately, many overloads of this type cannot be detected by overcurrent devices in the d-c leads to the tube. For example, in klystron amplifiers, an excessive r-f cavity voltage cannot be discerned from any of the r-f or d-c current readings. Yet, an excessively high cavity voltage may cause excessive voltage stresses on the ceramic walls of the klystron, which may cause it to crack. To prevent this, circuits of this type require an r-f sampling probe and circuit similar to that shown in Fig. 9-26. The rectified output of the r-f sampling probe will actuate a relay which will deenergize the transmitter if the r-f sampled voltage becomes too high.

Another similar condition exists in waveguide transmission line. If an arc occurs in the waveguide, the arc will travel down the waveguide transmission line toward the final amplifier. This arc must be detected before it hits the wall of the tube to prevent the tube wall from cracking. A common method of detecting this arc is through using a photocell

"looking up" the transmission line. This output of this cell actuates a relay interlocked with the transmitter control circuit in the event a waveguide arc occurs. This method is shown in basic form in Fig. 9-27.

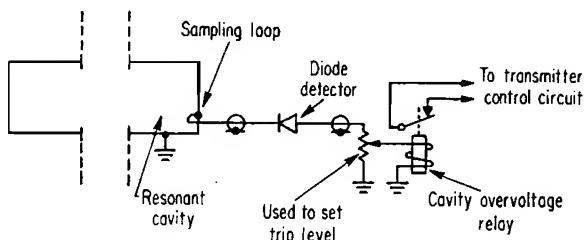


FIG. 9-26. Basic r-f overload-sensing circuit.

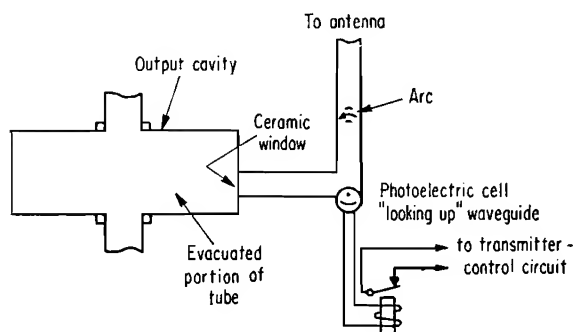


FIG. 9-27. Basic waveguide r-f arc-sensing circuit.

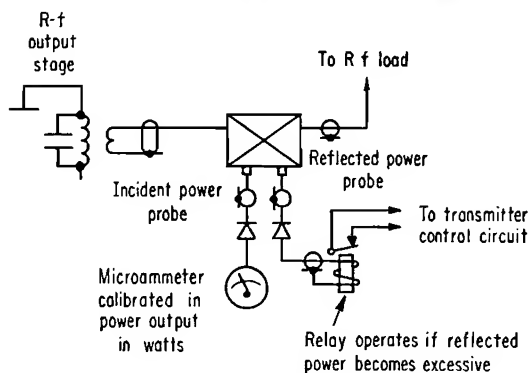


FIG. 9-28. Basic standing-wave ratio sensing circuit.

The transmission line as well as the tubes supplying the transmission line are protected against excessive standing-wave ratio by a detector sampling the reflected r-f power of the transmission line. This reflected power is sampled by means of a directional coupler, the principles of which are discussed in Chap. 11. This is shown in Fig. 9-28.

9-17. Across-the-line Diagrams. Complete control and protective circuit diagrams are often the most complicated and the least obvious in the transmitter. In an effort to make a logical presentation of the various functions of the control circuit, across-the-line diagrams or ladder diagrams are often used. In this type of diagram, the control lines are separated from the other functions of the transmitter, and the contacts of relays and circuit breakers are separated from their operating coils. A sample of this technique is shown in Fig. 9-29. The control voltage is

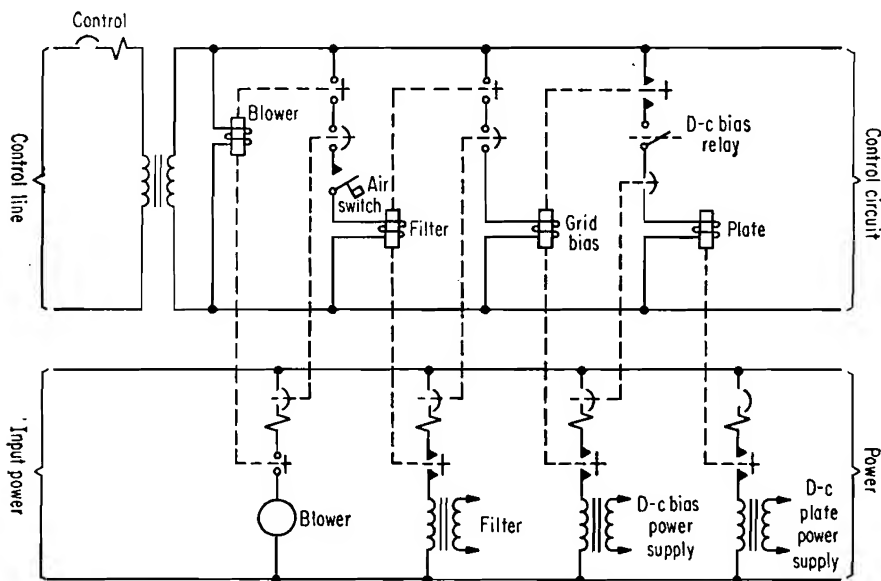


FIG. 9-29. Typical across-the-line diagram.

shown as two parallel lines. The necessary switching is shown as taking place across these two lines. The relationship of a coil and its contacts is maintained by identifying the coil and contacts of the same relay or circuit breaker with a letter or number code, or by connecting the coil and contacts with a dotted line. Rigid adherence to this type of diagram is often difficult. For this reason, many manufacturers of transmitting equipment make some modifications and variations on this method.

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COOLING

10-1. Transmitter Cooling. *General.* Transmitter cooling becomes a greater problem as the transmitter power input increases and as the size of the unit shrinks. Generally, in transmitting equipment, anywhere from 20 to 80 per cent of the input power is dissipated as heat in the tubes and other components. The resulting temperature rise must be held within tolerable limits to ensure satisfactory life for both the tube and other components in the equipment. Since the final amplifier stages normally have the greatest power input, they will also develop the greatest amount of heat. Therefore, most of the cooling effort in a transmitter is concentrated on the higher-power tubes. This heat can be removed in any of three ways: (1) through normal air-convection currents, (2) through forced-air cooling, or (3) through liquid cooling.

10-2. Forced-air Cooling of Tubes. The volume of airflow required to keep the anode temperature within tolerable limits is given in the tube data for every tube requiring forced-air cooling. In addition, the air pressure across the tube radiator to deliver this volume of air is also given. This pressure is given in inches of water. This is the most convenient unit for expressing and measuring the air pressure required for these applications.

Air-system Considerations. A typical forced-air system for a power tube is illustrated in Fig. 10-1. Basically an air system of this type is like an electrical series circuit in which each component in the circuit has a voltage drop depending on its resistance. The sum of these voltage drops must add up to the generator or battery voltage. The voltage drops in the series circuit are analogous to the pressure drops in an air system. Each duct, elbow, and filter in the air system of Fig. 10-1 has a pressure drop, and each of these pressure drops must add up to the total pressure available from the blower, just as in the series electrical circuit.

However, there is one significant departure of the air system from the electrical analogy just described. The pressure drop does not vary linearly with the rate of airflow. Instead, the pressure drop across any component in an air system will vary approximately as the square of the airflow rate. For example, if the airflow requirements were doubled, then

approximately four times the air pressure will be required to accomplish this increased flow. The pressure drops due to each component in the air system for different rates of airflow are available from standard tables for straight ducts, elbows, etc. The characteristics of the more special air-system components are available from the manufacturer of the item.

In order to select a blower, it is first necessary to determine the total pressure drop in the air system at the required rate of airflow. The

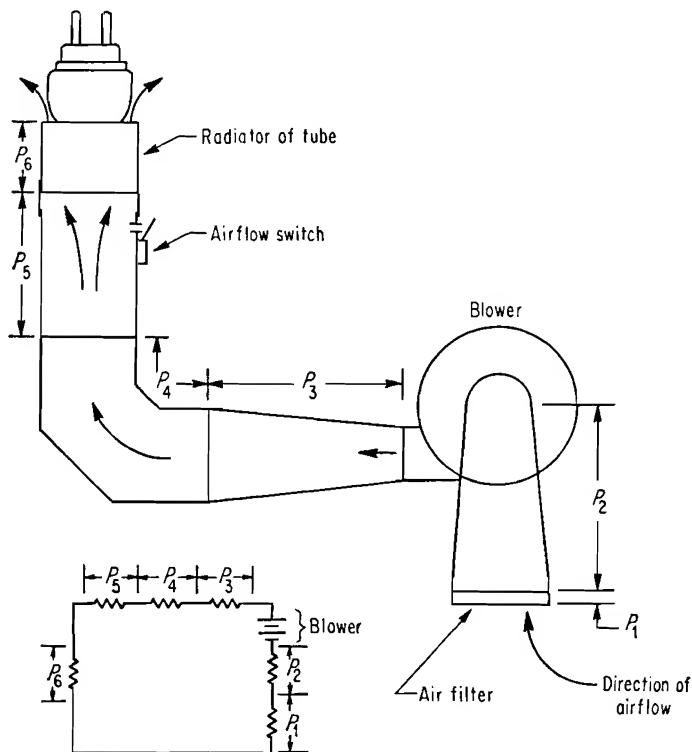


FIG. 10-1. A typical air system for a transmitting tube and equivalent electrical circuit.

blower selected must be capable of supplying this pressure at the required rate of airflow. The choice of a blower is made from the characteristics and data of blowers available from the numerous blower manufacturers.

The choice of a blower is sometimes modified by the fact that the duct supplying the tube does not have the same cross-sectional area as the blower outlet. This change in the cross-sectional area of the duct introduces an additional pressure drop which must be considered in the choice of a blower. This pressure drop must be added to the total of the other pressure drops in the system to determine the pressure requirements for the blower.

The pressure drop due to the change in cross-sectional area of the duct can be divided into two separate losses for ease in calculating the pressure drop. The first loss is that due to the change in the air velocity; the second loss is that due to air friction. The pressure drop due to the air velocity change is positive for a contraction in duct area, and negative for an expansion in duct cross-sectional area. This pressure drop is given by the equation

$$P_v = \frac{V_2^2 - V_1^2}{4,000^2} \quad (10-1)$$

where P_v = pressure drop in a duct due to air velocity changes, in. of water

V_1 = air velocity before change in area, ft per min

V_2 = air velocity after change in area, ft per min

These velocities can be calculated from the relationship

$$V = \frac{\text{cfm}}{A} \quad (10-2)$$

where V = duct air velocity, ft per min

cfm = airflow, cu ft per min

A = duct area at point of measurement, sq ft

The pressure drop due to the friction losses can be ignored if the change in the cross-sectional area is made over a length more than six times the larger duct diameter. If the change is made abruptly, i.e., in less than one diameter, then the air friction loss must be added to the duct losses in the system. This loss is positive for both an expansion or a contraction in duct cross-sectional area. An *expansion* in the duct area in the direction of the airflow will increase the pressure requirements of the blower according to the equation

$$P_{f1} = \frac{(V_1 - V_2)^2}{4,000^2} \quad (10-3)$$

where P_{f1} is the pressure drop in inches of water due to friction in an expanding air duct. The other factors are the same as those in Eq. (10-1).

A *contraction* of the duct area increases the pressure requirements of the blower by the amount

$$P_{f2} = \frac{K_c V_2^2}{4,000^2} \quad (10-4)$$

where P_{f2} = pressure drop due to friction in contracting air duct, in. of water

K_c = constant which depends on duct contraction and is determined from curve of Fig. 10-2

V_2 = air velocity after change in area, ft per min

Where the change in the cross-sectional area of the duct takes place in a length between one and six times the duct diameter, then the pressure drop can be obtained from Eqs. (10-1) through (10-4).

With the pressure drops and airflow requirements of the air system known, a curve known as the *system characteristic* can be determined and plotted on the blower characteristic curve. The blower characteristic curve is obtained from the blower manufacturer's literature and is the graphical plot of the blower outlet pressure in inches of water vs. the airflow of the blower in cubic feet per minute. The system-characteristic curve can be determined from the following equation which is then plotted for various values of the blower pressure vs. airflow curve.

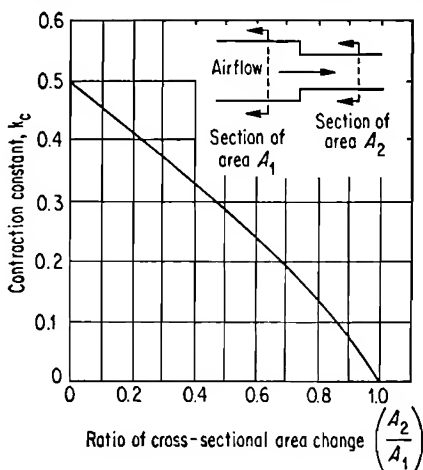


FIG. 10-2. Chart for determining contraction constant k_c for use in Eq. (10-4). (Courtesy of Electronics.)

$$(P_s)_x = P_s \left(\frac{Q_x}{Q} \right)^2 \quad (10-5)$$

where $(P_s)_x$ = unknown static pressure required by system, in. of water
 P_s = known static pressure required by system, in. of water
 Q_x = unknown value of airflow, cu ft per min
 Q = known value of airflow, cu ft per min

A typical blower and system-characteristic curve is shown in Fig. 10-3. The intersection of the system characteristic with the blower curve will give the pressure and delivery that will be present at the blower outlet for this particular system. The choice of the blower should then be checked against the original requirements to make sure the delivery and pressure is adequate. The pressure and delivery should exceed the minimum requirements to allow for excessively dirty filters, partial clogging of the tube radiator, air-system leaks, etc.

Blower Considerations for High-altitude Operation. Usually the tube manufacturer's data is based on the density of air at sea level. To obtain the equivalent cooling at different altitudes, it is necessary to maintain the same weight of air being forced through the tube radiator per minute. As the altitude increases, the density of the air decreases, i.e., the weight per cubic foot decreases. This means that it is necessary to increase the volume of air at the higher altitudes to obtain the equivalent sea-level

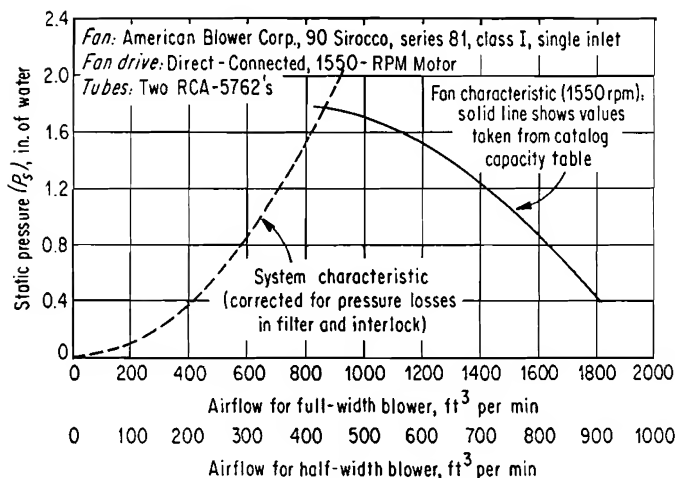


FIG. 10-3. A typical blower and air system characteristic curve. The intersection of the two curves gives the pressure and delivery of the blower. (Courtesy of Electronics.)

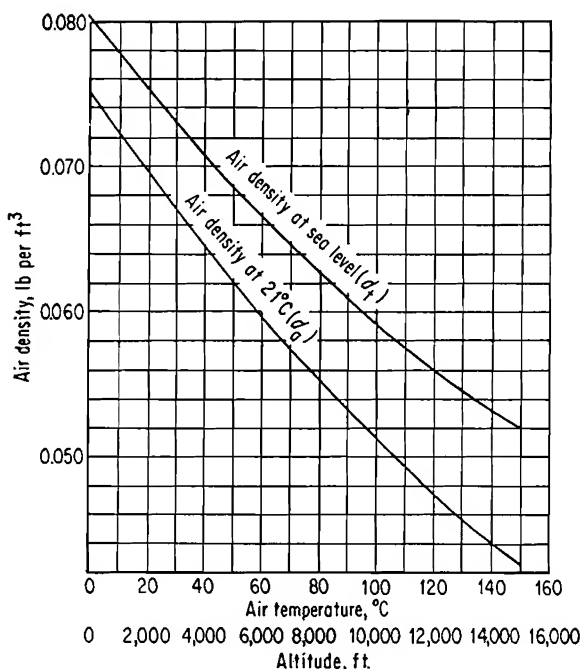


FIG. 10-4. Change in air density at increased air temperatures and altitudes. (Courtesy of Electronics.)

cooling. The density of the air at the higher altitudes is found in the curve of Fig. 10-4. The blower requirements are then determined from the following equations.

$$Q_2 = \frac{0.075Q_1}{D_2} \quad (10-6)$$

where Q_2 = airflow at high altitude, cu ft per min

Q_1 = airflow required at sea level, cu ft per min

D_2 = air density at high altitude, lb per cu ft

$$(P_s)_2 = \frac{0.075(P_s)_1}{D_2} \quad (10-7)$$

where $(P_s)_2$ = pressure at high altitude, in. of water

$(P_s)_1$ = pressure at sea level, in. of water

D_2 = air density at high altitude, lb per cu ft

This will give the pressure and delivery in cubic feet per minute required at the higher altitude. However, in order to specify the blower, it is still necessary to determine what a given blower will deliver at the normal sea-level conditions, since this is the conditions at which the blower would be rated. A blower can be regarded as a constant-delivery device; i.e., it will deliver the same cubic feet per minute at the higher altitudes as at sea level. However, the required sea-level pressure delivered at the outlet of the blower will change from the value calculated in Eq. (10-7). This can be determined from the relationship

$$(P_s)_1' = \frac{(Q_2)^2(P_s)_1}{(Q_1)^2} \quad (10-8)$$

where $(P_s)_1$ = blower outlet pressure at sea level with delivery of Q_2

Q_1 = airflow required for tube at sea level, cu ft per min

Q_2 = airflow at high altitude, cu ft per min, calculated from Eq. (10-6)

$(P_s)_1$ = pressure originally required at blower outlet at sea level

Blower Considerations for High-temperature Operation. A problem similar to that discussed above exists when cooling tubes with ambient air temperatures above normal. The normal temperature is that temperature at which the manufacturer of the tube has specified the air-pressure and air-temperature ratings of the tube. This temperature varies with both the tube type and the manufacturer.

The curve of Fig. 10-4 gives the change in the air density at increased air temperatures. The blower requirements are determined in the same way as for the high-altitude case.

Tube-outlet Air Temperature. The temperature of the tube outlet air is of interest to ensure that any components in this airstream are

adequately rated for this temperature. This temperature can be determined from the equation

$$T_o = \frac{(T_i + 273)W_d}{164Q} + T_i \quad (10-9)$$

where T_o = outgoing temperature, °C

T_i = incoming temperature, °C

W_d = dissipation, watts

Q = airflow, cu ft per min

This equation can also be worked backwards to find the plate dissipation and thus determine the power output from the amplifier.

Measurement of Air Pressure. Air pressure in air-cooled systems is commonly measured in inches of water. These measurements are easily

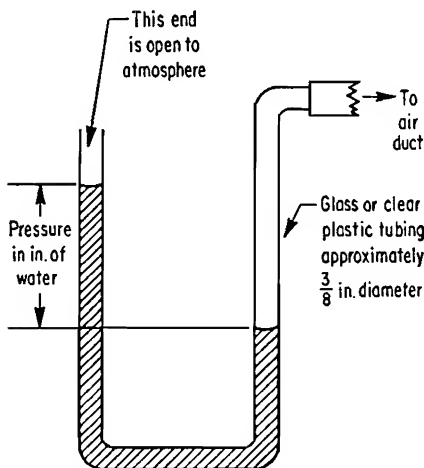


FIG. 10-5. A simple manometer for measuring air pressure in forced air systems.

made on any air system with a manometer; the latter instrument can be formed from a piece of glass or plastic tubing. A simple manometer for this purpose is shown in Fig. 10-5. The open end is left open to the atmosphere. The opposite end is connected to the air system by means of a piece of plastic or rubber hose. It is an absolute necessity that this connection be air-tight to prevent errors in the pressure reading. The connection to the air system is made through a hole in the duct. This should be at least 3 in. away from any change in duct size or from any other obstruction in the duct.

Measurement of Air Velocity. A velometer is used to measure air velocity in an air system. Two commonly used basic types of velometers are shown in Figs. 10-6 and 10-7. The mechanical type shown in Fig. 10-6 uses a sample of the air conducted through a rubber tube placed in the airstream. The air sample is passed over a vane to which is attached a pointer. The up-scale reading of the pointer over a calibrated scale will indicate the air velocity in feet per minute.

A second type, shown in Fig. 10-7, uses a self-heating thermistor in a Wheatstone-bridge circuit. If the thermistor element is placed in the airstream, the resistance value of the thermistor will change as a result of the cooling effect of the airstream on the thermistor. This will cause

the bridge to become unbalanced to a degree depending on the air velocity. This unbalance can be read on a meter calibrated in feet per minute of air velocity.

Air-system Protection. Air-system flow can be monitored and interlocked with the transmitter control circuits to prevent continued operation of the equipment in case of air-system failure. The usual method is to use an air switch that is actuated by the air-system pressure or velocity.

A pressure-operated switch may use a diaphragm or piston type of chamber to actuate a sensitive switch. The velocity-operated switches are operated by a vane. The size of the vane determines the velocity rating of the switch.

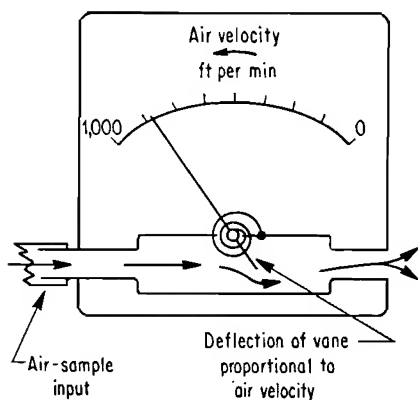


FIG. 10-6. Basic operation of a mechanical velometer for measuring velocity of an air stream.

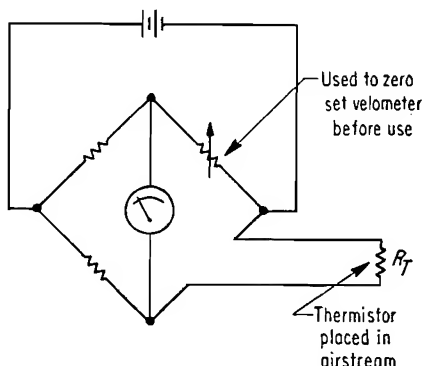


FIG. 10-7. Basic operation of an electrical velometer.

A switch actuated by air pressure may be placed in the air system between the blower outlet and the tube. It is in this region that the air pressure is sufficiently high to operate the switch. Some degree of care must be exercised in using this type of switch, to be sure that, if the air ducting breaks or cracks, the air pressure will drop sufficiently to operate the switch before damage can be done to the tube.

Velocity air switches are preferably placed in the tube air outlet. Thus, if the airflow through the tube is reduced for any reason, the air switch will operate. In many cases, the velocity type of air switch is used to monitor the air-system pressure by placing the air switch on the outside of the duct, over a small hole in the air duct. The velocity of the air escaping through the hole is proportional to the air pressure within the duct.

10-3. Liquid Cooling of Tubes. For amplifier tubes with plate dissipations of approximately 25 kw and up, the most satisfactory method of

removing the heat from the plate of the tube is by means of a circulating liquid coolant. A typical liquid-cooled system is shown in schematic form in Fig. 10-8.

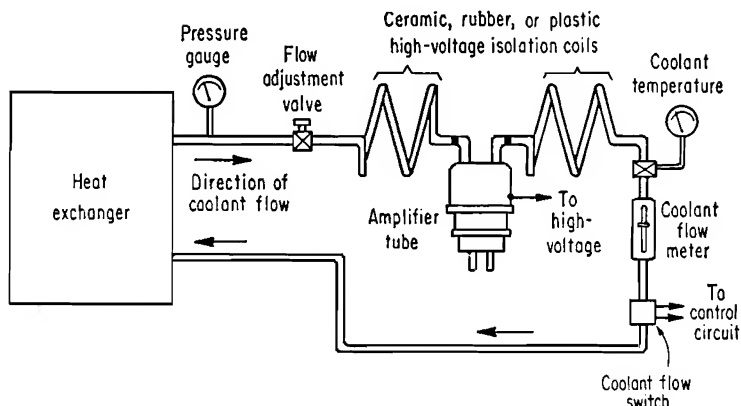


FIG. 10-8. A typical liquid-cooled system.

The circulated liquid is usually distilled water. Distilled water, aside from being economical, is also an electrical insulator. This makes it desirable in systems in which the coolant comes in contact with the high voltage on the plate of the tube. In equipment designed for extremely cold weather, other coolants are available. An ethylene glycol and water solution is useful in cases where the coolant need not be an insulator. Other coolants such as Monsanto OS-45 and Minnesota Mining FC-75 are occasionally used as coolants for cold-weather equipment. These latter two coolants are insulators; however the low specific heat of these liquids requires the use of greater flow rates and higher pressures in the system.

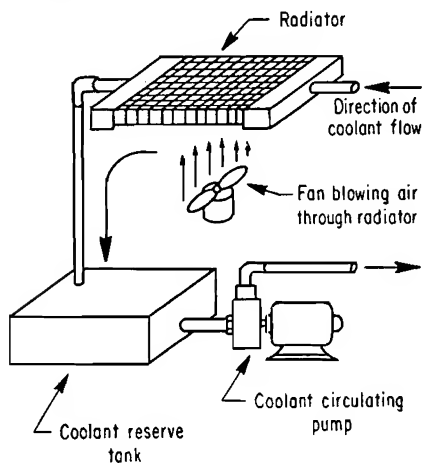


FIG. 10-9. Basic elements of a liquid-to-air heat exchanger.

heat is transferred to the air, the unit is known as a *liquid-to-air heat exchanger* (shown in Fig. 10-9). If the heat is transferred to another liquid, the unit is known as a *liquid-to-liquid heat exchanger* (shown in Fig. 10-10).

The pump for either type of heat exchanger must be capable of supplying the required rate of coolant flow. Therefore the pump must develop a sufficient pressure to overcome all the pressure drops in the system. The pressure drops at the rate of flow required can be estimated from standard tables of pressure drops for piping and pipe fittings. The manufacturer's literature usually lists the necessary data to obtain the pressure drops in flowmeters, isolation coils, flow switches, and the tube.

In addition, the same considerations apply to the air system of the liquid-to-air heat exchanger as those previously outlined for the direct forced-air cooling of tubes.

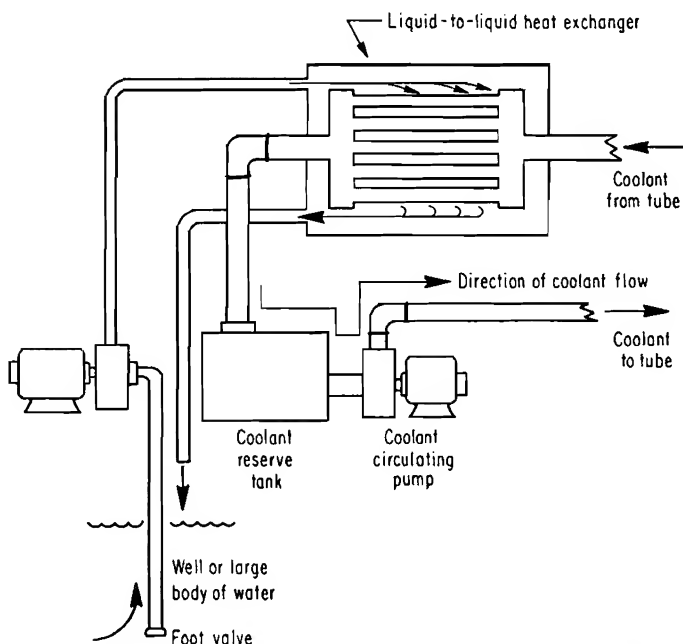


FIG. 10-10. Basic elements of a liquid-to-liquid heat exchanger.

Maintenance of a Liquid-cooling System. Liquid-cooling systems generally require more maintenance than the equivalent air-cooling system. This is due in part to the additional mechanical parts involved in a liquid-cooling system, in part to the corrosive effect of the coolant, and in part to the electrolytic action resulting from different parts of the liquid-cooling system operating at different voltages. In addition, scale formation must be minimized to ensure proper heat transfer between the tube and the liquid coolant.

The damaging effects of corrosion and scale can be minimized by using a coolant of either distilled water or deionized water. The water used in the cooling system should have a minimum initial resistance of

100,000 ohms per cu cm. The coolant should be replaced when the coolant resistance falls to 20,000 ohms per cu cm. This can be calculated from the leakage current, the voltage across the liquid-cooling system (through which the leakage occurs), the total cross-sectional area of the liquid-coolant supply pipes to the tube, and the length of the leakage path. The area required for this calculation is the sum of both the inlet and outlet pipes to the tube. Electrically, the leakage resistance offered by the coolant in these two pipes is in parallel.

In liquid-cooling systems using an ethylene glycol-water mixture, the use of rust and corrosion inhibitor is usually recommended. This type of coolant is an excellent electrical conductor and is used in systems where the anode is very close to ground potential. This is commonly the case in klystron power amplifiers.

Removing Scale Formation. Most scale will form on the heated anode surface. However, it is not generally recommended that the tube be removed for such maintenance routines. Excessive handling of the tube may only invite an early tube failure. As filaments are used, they tend to become very brittle. The shock of removing the tube may cause the filament strands to snap. Scale removal can be accomplished while the tube is in place by adding trisodium phosphate to the coolant. This is circulated through the system for approximately 1 hr. The water temperature should be around 140°F. At the end of this time, drain the system completely. This may be followed by a detergent flush. Any ordinary household detergent is satisfactory. This should be drained out and followed by a water flush. The water may be run into the system continuously with the drain valve for the system open. A final distilled-water flush is generally advisable, after which the system may be refilled for use. Any filters in the system should be removed during this process to prevent clogging with the dislodged sludge and scale.

Tubes with external anodes with removable water jackets may be cleaned by removing the outer jacket and cleaning the anode surface with a 20 per cent solution of hydrochloric acid and a cloth.

Protection of a Liquid-coolant System. Cooling systems require electrical interlocking with the control circuit of the transmitter to protect the tube from damage due to a coolant-system failure. These failures may be adequately protected against through the use of a temperature switch and coolant-flow switches in the coolant outlet line from the tube. These switches can be arranged to remove the voltages from the tube in the event that the coolant temperature becomes excessive, or that the coolant flow drops below the minimum required for safe operation.

In addition, the coolant pressure present at the anode of the tube should not exceed the tube manufacturer's recommendation. If a stop-

page or obstruction occurs, the pressure at the tube anode may exceed the maximum recommended for the tube. A pressure relief valve may be installed in the inlet coolant line to the tube to protect against this fault. It should be installed as close to the tube as is practicable for maximum protection.

10-4. Tube Temperature Measurement. The basic purpose of any cooling system is to limit the bulb and seal temperature to the values recommended by the tube manufacturer. Thus, a check on the effectiveness of any air- or liquid-cooling system is to measure these temperatures and compare them to the maximum values recommended for the tube.

A convenient method of doing this is by using temperature-sensitive paints which will melt at a particular temperature. By using an assortment of these paints, each of which will melt at a different temperature, the bulb and seal temperatures can be accurately determined.

10-5. General Cabinet Cooling. Transformers, rectifiers, resistors, and the blowers themselves all contribute some amount of heat. If this heat is confined within the cabinet, the resultant temperature rise may be more than the components can safely stand. The airflow necessary to remove this heat and hold the temperature rise to that which the components in the transmitter can safely stand is found by the equation

$$Q_a = \frac{W}{C_p T_r} \quad (10-10)$$

where Q_a = airflow, cu ft per min

W = electricity dissipated as heat, watts

T_r = temperature rise of air, °C

C_p = specific heat of air (For dry air at normal atmospheric pressure, this is usually 0.61.)

Equation (10-10) means that 1 watt of power dissipated as heat will produce a temperature rise of 0.61°C if the airstream is flowing at 1 cu ft per min.

Thus, if the maximum or desired temperature rise in the transmitter cabinet is known, and if the electricity in watts dissipated as heat from the miscellaneous components within the cabinet is also known, then the volume of air necessary to hold this temperature rise can then be determined.

General cabinet cooling is generally achieved with a propeller-type fan which pulls air into the cabinet rather than sucking air from the cabinet. In this way, the cabinet is under some small amount of pressure which prevents unfiltered air from being sucked into various openings and cracks in the cabinet.

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R-F COMPONENTS

11-1. R-F Components. General. Apart from the more commonly used components in the transmitter, most of which are readily understood operationally and functionally, there is a group of r-f components whose operation and application are not completely obvious at a superficial glance. The purpose of this chapter is to discuss the principles and applications of the more commonly used r-f devices of this type.

11-2. Directional Couplers. General. A directional coupler is a device for individually sampling the r-f incident and reflected wave in a transmission line. The operation of any transmission line can be described in terms of the r-f incident and reflected waves. The incident wave on a transmission line results from the propagation of the generated r-f signal toward the load. The reflected wave results from any mismatch, i.e., if the transmission line is terminated in a resistor value equal to anything other than the characteristic impedance of the transmission line. The addition of the incident and reflected waves at any instant result in the familiar standing-wave voltage and current distribution on the transmission line.

The relationship between the incident and reflected power sampled by the directional coupler and the standing-wave ratio present on the transmission line is expressed by the equation

$$K = \frac{1 + R}{1 - R} \quad (11-1)$$

where K = standing-wave ratio

R = reflection coefficient or ratio of relative amplitude of reflected signal to incident signal

Application. The directional coupler is widely used (in conjunction with an auxiliary meter and detector) to determine the standing-wave ratio present on a transmission line. This transmission line may be between stages in a transmitter or between the transmitter and the antenna. A method of using the directional coupler for standing-wave-ratio measurements is shown in Fig. 11-1. The meter is normally adjusted for a full-scale reading with the switch in the incident position.

When the switch is in the reflected position, the meter will indicate the standing-wave ratio present on the transmission line. The meter scale is specially calibrated in standing-wave ratios.

The reflected signal sampled by the reflected probe of the directional coupler is often used to energize a protective relay in the transmitter control circuit whenever the reflected power exceeds a preset level. An excessively high reflected signal is indicative of a mismatch and a potentially dangerous condition. A typical application for this is illustrated in Sec. 9-16.

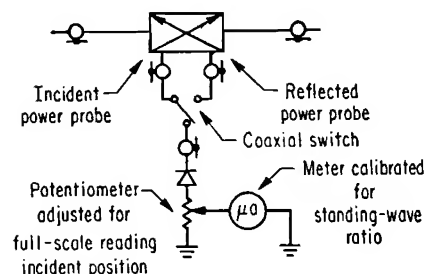


FIG. 11-1. Use of the directional coupler for measuring standing-wave ratio.

Ignoring the effect of the reflected wave on the power output will introduce an error which becomes appreciable only if the reflected signal is large. For example, if the standing-wave ratio is 2:1, an error of only 10 per cent will result in the power reading taken from the incident probe of the directional coupler.

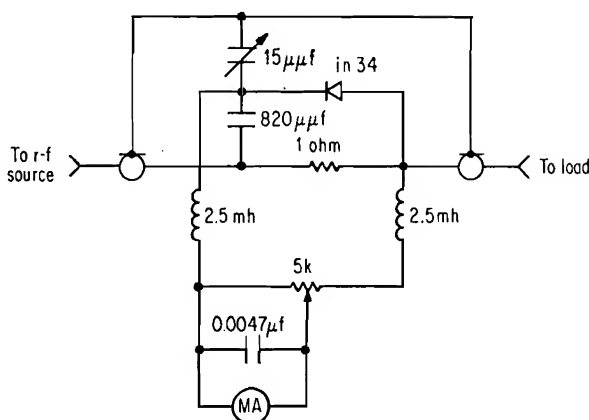


FIG. 11-2. Resistance directional coupler.

Operation. Many ingenious types of directional couplers have been devised since their inception in the early 1940s. However, the discussion that follows will be limited to the representative types of directional couplers more commonly found in radio transmitters.

The Resistance Directional Coupler. A typical directional coupler of this type is shown in schematic form in Fig. 11-2. This directional

coupler can be simplified to that shown in Fig. 11-3, where it can be recognized as the familiar bridge circuit. From this, one can intuitively see that, once the bridge is balanced, any change in the value of the terminating impedance Z will result in an unbalancing of the bridge and cause a reading on the meter V . However, a better and more complete description of the operation of the directional coupler can be given with the aid of some simple mathematics.

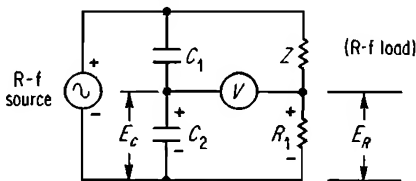


FIG. 11-3. Simplified form of the directional coupler of Fig. 11-2.

The voltage and current on any transmission line can be expressed as

$$E = K_1 E_i + K_2 E_r \quad (11-2)$$

$$I = \frac{1}{Z_0} (K_1 E_i - K_2 E_r) \quad (11-3)$$

where E = rms voltage at any point on transmission line

E_i = incident rms voltage

E_r = reflected rms voltage

K_1 and K_2 = factors related to phase of the respective voltages and can be considered to be one

I = rms current at any point on transmission line

Z_0 = surge impedance of transmission line

The voltage across the capacitor C_2 can be expressed as

$$E_c = \frac{C_1}{C_2} E = \frac{C_1}{C_2} (K_1 E_i + K_2 E_r) \quad (11-4)$$

where the factors are the same as described previously for Eqs. (11-2) and (11-3) or in Fig. 11-3. The voltage across the resistor sampling the transmission-line current is

$$E_r = R_1 I = \frac{R_1}{Z_0} (K_1 E_i - K_2 E_r) \quad (11-5)$$

where the factors are the same as described previously for Eqs. (11-2) and (11-3) or in Fig. 11-3.

If the polarities of the voltages due to the generator are assigned as shown in the simplified schematic of Fig. 11-3, it becomes apparent that the meter is reading the difference of the voltages of Eqs. (11-4) and (11-5). This results in the equation

$$V_1 = \frac{2R_1}{Z_0} E_r \quad (11-6)$$

where V_1 is reading of voltmeter V of Fig. 11-3. The remaining factors have been described for Eqs. (11-2) and (11-3) and in Fig. 11-3.

This plainly states that the meter reading is only due to the reflected voltage for the circuit of Figs. 11-2 and 11-3. This voltage is ideally zero when the transmission line is terminated in its characteristic impedance.

If the directional coupler is reversed, i.e., the positions of the generator and terminating impedance are reversed, the polarities across the resistor R_1 sampling the transmission-line current will also reverse. This will alter the signs of Eq. (11-5) and will result in the following equation

$$V_2 = \frac{2R_1}{Z_0} E_i \quad (11-7)$$

where V_2 is the reading of voltmeter V of Fig. 11-3. The remaining factors have been described for Eqs. (11-2) and (11-3) and in Fig. 11-3.

So, it is clear that in this case the meter will respond only to the incident wave.

Loop Type of Directional Coupler. A cross section of the loop type of directional coupler is shown in Fig. 11-4. This probe is coupled to the

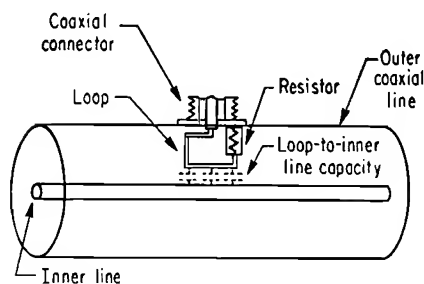


FIG. 11-4. Loop type of directional coupler.

inner line of the transmission line both capacitively and inductively. For a wave traveling to the right, the capacitance of the probe to the inner line will result in a current flowing in the probe as shown in Fig. 11-5a. For the same wave, the inductive coupling of the probe will induce a current flow in the probe as shown in Fig. 11-5b. In this case, the capacitive and inductively coupled currents will add at

the r-f output probe and will subtract for those flowing through the resistive termination.

The probe output voltage will be proportional to either an incident or a reflected wave according to the orientation of the probe with respect to the transmission line. By rotating the entire directional-coupler probe assembly, the directional coupler will read the reverse condition; i.e., if the probe output was proportional to the incident wave, then reversing the probe assembly will cause the output to be proportional to the reflected wave. Many directional couplers are made in just this way. This permits one directional-coupler probe to sample both the incident and reflected waves with precisely the same degree of coupling in both cases.

Normally, the directional-coupler capacitive and inductive probe currents are made equal by proportioning the loop. If a wave is present in only one direction as would be the case for a perfectly matched load, then no current would flow through the resistive termination for the probe. However, if the r-f load is not correct, a reflected wave will result and will travel toward the left in Fig. 11-5. This will result in

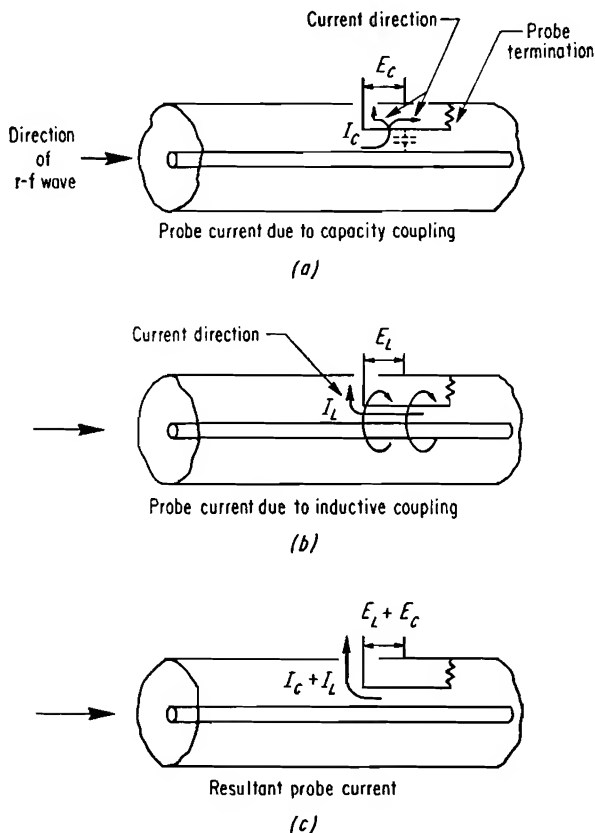


FIG. 11-5. Coupling present in loop type of directional coupler.

current flowing through the probe-terminating resistor. Thus the probe-terminating resistor must be capable of dissipating power according to the amount of coupling between the probe and the transmission line and according to the power level in the transmission line. For example, if the maximum power in the transmission line is 1,000 watts and the directional coupler probe has 30 db of coupling, then the power dissipated by this resistor will be 1.0 watt.

At the higher-power levels, placing a resistor of this size inside a transmission line presents a problem both mechanically and electrically. In

these cases, the probe can be constructed with the termination outside (shown in Fig. 11-6). This arrangement also permits the probe to be used for both incident and reflected power measurements simply by switching the terminating load from one probe to the other by means of a coaxial switch or relay.

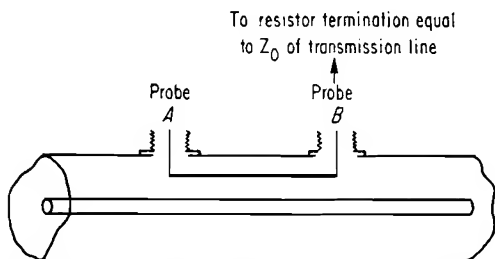


FIG. 11-6. Coupling in loop type of directional coupler with external load.

The use of separate probes for the incident and reflected waves permits the coupling for each probe to be individually adjusted. This is desirable in cases where the standing-wave ratio is low and consequently the output of the reflected probe is also very low. Increasing the coupling of the reflected probe will increase the output from the probe and make it easier to meter.

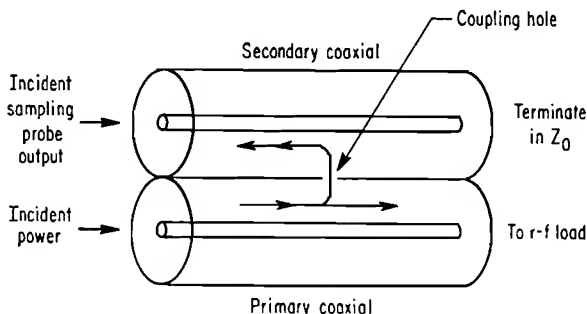


FIG. 11-7. Single-hole directional coupler.

The Single-hole Directional Coupler. This type of directional coupler uses a hole in the outer conductor of a coaxial or waveguide to provide the necessary electric and magnetic coupling. A typical single-hole directional coupler is shown in Fig. 11-7. The operation of this type of directional coupler is otherwise identical with that described for the loop type of directional coupler described previously in this section. The method by which the hole provides the electric and magnetic coupling is illustrated in Fig. 11-8. The electric coupling can be adjusted by

varying the length of the hole in the direction of the inner line of the coaxial line. The magnetic coupling can be adjusted by making the slot wider in a direction at right angles to the inner conductor. The directional characteristics of the directional coupler are achieved when the magnetic and electric coupling are equal. In this type of directional coupler, the secondary line shown in Fig. 11-7 must be accurately terminated in a value equal to the surge impedance of the transmission line.

If the coupling hole is circular, then the magnetic and electric coupling between the primary and secondary transmission lines can be adjusted by skewing the position of the secondary with respect to the primary.

By switching the termination of the secondary coaxial line from one end to the other, the directional coupler can also be used to sample both the incident and reflected power.

The Two-element Directional Coupler. In this type of directional coupler, the means of coupling to the primary transmission line is spaced an odd multiple of a quarter wavelength at the operating frequency. The coupling must be either electric or magnetic for both probes. Thus the coupling probes may be both loops (magnetic coupling) or capacitive probes (electric coupling) or specially shaped holes which produce only electric or magnetic coupling.

The two-element directional coupler is shown in schematic form in Fig. 11-9. Magnetic coupling is used for the purposes of this example. Other forms of coupling can be used. The two-element directional coupler is usable with both coaxial and waveguide transmission lines. In Fig. 11-10, a wave traveling down the line in the direction shown will set up two waves in the secondary transmission line as a result of the energy picked up at the sampling loop at *A*. Similarly the loop at *B* will also set up two waves traveling in the two opposite directions as shown. However, since the waves traveling the path *ACD* and *ABD* both travel the same distance, they will add or reinforce each other.

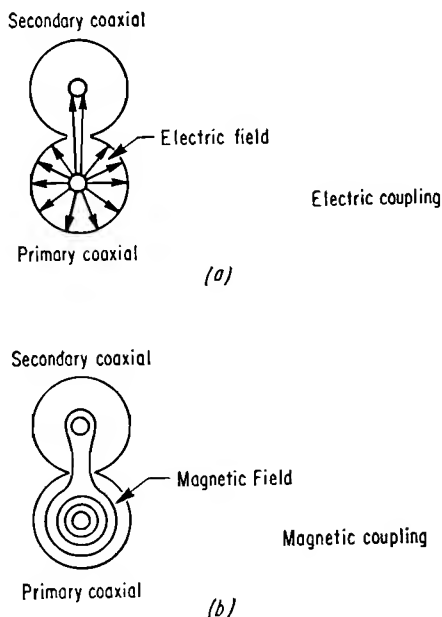


FIG. 11-8. The electric and magnetic coupling in the single-hole directional coupler.

However, the waves traveling the path AC and $ABDC$ are different by a half wavelength, which will cause them to subtract or cancel each other. The net result is that the forward or incident wave produces a signal in

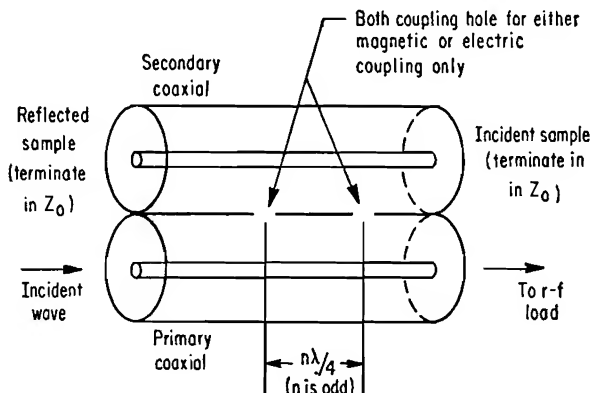


FIG. 11-9. The two-element directional coupler.

the secondary transmission line only in the forward direction. Conversely, the reflected wave will produce a signal in the secondary line in the opposite direction only.

11-3. Dummy Loads. Dummy loads are used to provide a nonradiating means of dissipating r-f energy. As such, dummy loads are used to

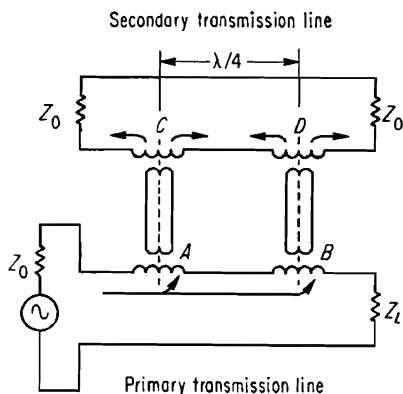


FIG. 11-10. Electrical equivalent of the two-hole directional coupler.

dielectric of a transmission line. In addition, dummy loads can be physically divided into types requiring liquid cooling and those requiring air cooling.

Resistance Dummy Loads. At the low and medium frequencies, the resistance load can be a relatively simple affair made up of multiple

dissipate the output of transmitters for testing purposes and, in other cases, are used to provide resistance loading for tuned circuits for broadband applications. Dummy loads are also used in r-f filters to dissipate the transmitter output power over the frequency band rejected by the filter.

Electrically, dummy loads can be divided into the resistance type in which the r-f power is dissipated in a resistor, and the lossy dielectric type in which the r-f power is dissipated in the intentionally lossy

combinations of either noninductive wire or carbon resistors. R-f match at these frequencies is generally no problem since the wavelength compared to the resistor length is quite small.

In the cases where the match is critical, i.e., where the inductance of the load is excessive, it can be adjusted by a series capacitor as shown in Fig. 11-11.

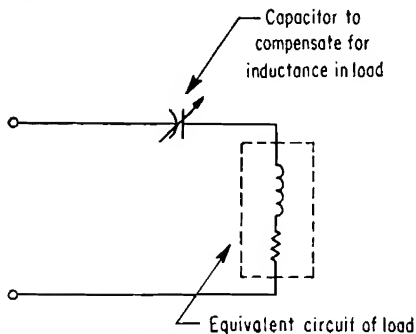


FIG. 11-11. Low-frequency dummy load.

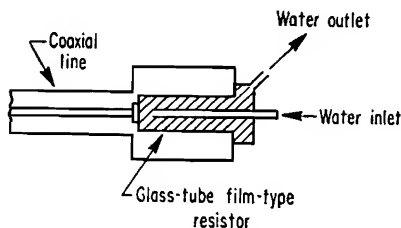


FIG. 11-12. Water-cooled vhf dummy load.

However, adjusting or tuning the load by these methods also restricts the frequency band over which the load will perform satisfactorily. If the load is constructed of matched transmission-line sections as shown in Fig. 11-12, the inherent frequency range of the dummy load will be greatly increased. In loads of this type, water is often circulated across the resistance film to carry away the heat developed in the resistor. The

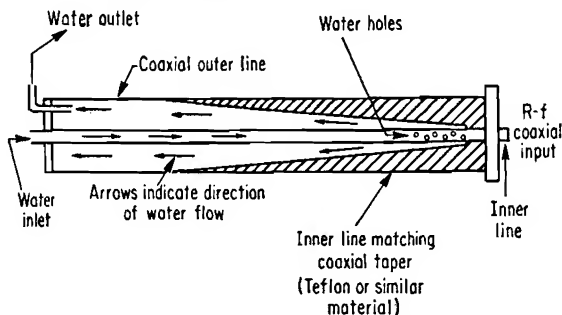


FIG. 11-13. Lossy dielectric dummy load.

water supply for the dummy load is usually connected to the transmitter heat exchanger. This heat exchanger must have the added capacity to dissipate the entire power output as well as the dissipation losses of the transmitter.

11-4. Lossy Dielectric Loads. This type of load usually uses the distilled water to serve as the lossy dielectric for a section of coaxial line. The basic construction of such a load is shown in Fig. 11-13. It is

apparent from the construction of such a load that the d-c resistance across the input terminals is essentially zero. This is not true for the resistance type previously described. The tapered section shown in Fig. 11-13 serves to match the air-dielectric transmission line to the transmission-line section using the lossy dielectric liquid. As before, the lossy dielectric i.e., distilled water, is circulated through a common heat exchanger for the transmitter.

11-5. R-F Filters. Radio-frequency filters are often used in transmitting equipment to restrict or limit the transmitter output frequencies. As such, the filter must be constructed to handle the high voltages and currents that may develop. To limit the power-handling requirements of the filter, the filter is sometimes located in one of the earlier stages of the transmitter and then followed by linear amplifier stages. This technique is shown in Fig. 11-14. This arrange-

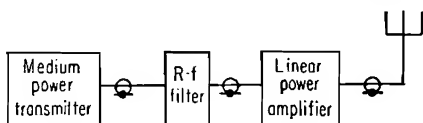


FIG. 11-14. A r-f filter may be placed in the low power stages if followed by linear amplifier stages.

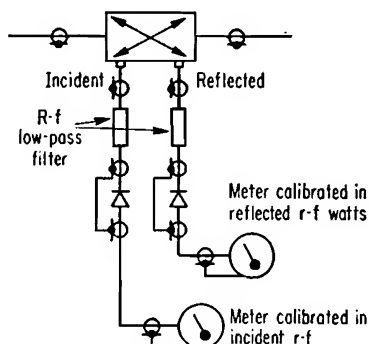


FIG. 11-15. Low-pass r-f filters are often used in conjunction with directional couplers to provide accurate power readings.

ment is feasible providing the amplifier stages following the filter are all perfectly linear. Any nonlinearity in the amplifier stages following the filter will result in some spurious output.

Low-power filters are often used between the directional-coupler probe and the crystal detector operating the power meter (see Fig. 11-15). This prevents the transmitter harmonics from affecting the power meter reading, i.e., the power meter will read the fundamental frequency power only.

The electrical design of a filter is independent of the operating frequency. Thus the filter design equations listed in the appendix are equally applicable for all filters. Greater attenuation characteristics can be achieved simply by cascading the appropriate sections. However, the pass-band losses will also increase when this is done.

The mechanical design of a filter will vary with frequency. As the frequency of operation increases, the lead length and the choice of components become more important. Similarly, as the attenuation characteristics of the filter increase, the shielding and isolation between filter components must also increase. At the vhf and uhf frequencies, the

lumped circuit components, such as capacitors and inductors, are replaced by the equivalent coaxial transmission-line sections which have been previously described in Sec. 5-16.

11-6. Vestigial-sideband Filter. The vestigial-sideband filter is commonly used in television applications to eliminate the lower sideband in accordance with the FCC requirements. The vestigial-sideband filter may be of either the constant-resistance type or the pure-reactance type. The constant-resistance filter is the most commonly used type since it provides a constant resistance to the transmitter final output stage. The constant-resistance filter may consist of a low-pass filter and a high-pass filter connected in parallel as shown in the schematic of Fig. 11-16. In this case, the high-pass filter will reject the lower sideband, and the lower-sideband energy will be dissipated in the dummy load connected to the

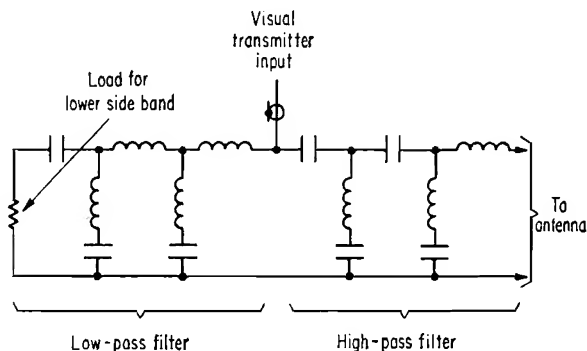


FIG. 11-16. A constant-resistance vestigial-sideband filter.

output of the low-pass filter. Conversely, the upper sideband will not pass through the low-pass filter but will pass through the high-pass filter to the antenna.

The second type of filter that can be used for this purpose is the pure reactance type of filter made up of conventional constant K and M derived filter sections. This type does not provide a constant resistance load to the transmitter. However, it is simpler and less costly to construct. At the rejection frequencies, this type of filter will reflect the energy back toward the generator or transmitter output stage. The reflected energy is absorbed or at least partially absorbed by the internal impedance of the output stage. To minimize the effects of the reflected energy in the final picture, the transmission line connecting the transmitter and the vestigial-sideband filter should have a length corresponding to a maximum transmission time of $0.01 \mu\text{sec}$. At the vhf frequencies and for the usual air type of transmission line, this corresponds to a maximum length of 10 ft between the transmitter and the vestigial-sideband filter.

11-7. Diplexers. General. Diplexers are used to combine the outputs of two transmitters into a common load and at the same time provide isolation between the two transmitter outputs.

Diplexers are commonly employed in television transmitters where it is desired to have a common antenna for both the aural and visual transmitters. Diplexers are also used to combine the output of two or more transmitters operating at the same frequency to produce a greater power output. There are many types of diplexers which will fulfill the general requirements described above; however, the specific type used depends on (1) whether the transmitters to be combined are on the same frequency or whether the operating frequency of each transmitter is different, (2) whether the combined transmitter frequency is fixed, or whether the operating frequency may be changed during the course of operation, and (3) whether the antenna requires one or two separate output transmission lines from the diplexer. Some general types of diplexers described in the following material which will fulfill the specific

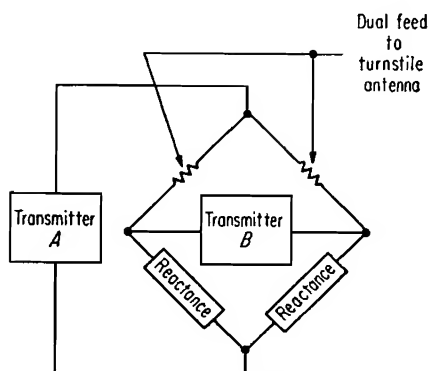


FIG. 11-17. A balanced bridge diplexer.

requirements are (1) the balanced-bridge type, (2) the slotted-bridge type, (3) the notching-filter type, (4) the magic T, (5) the coaxial-hybrid ring.

The Balanced-bridge Diplexer.

This type of diplexer has an equivalent circuit similar to a Wheatstone bridge (see Fig. 11-17). It is apparent from Fig. 11-17 that, if the bridge is properly balanced, none of the visual input will appear at the aural transmitter output, and vice versa. It can also be

noted from Fig. 11-17 that this type of diplexer requires two transmission-line feeders to the antenna. The commonly used super-turnstile type of television antenna does use two transmission-line feeders.

In the circuit shown in Fig. 11-17, the transmitter *B* is actually shunted by the reactance arms of the bridge. For this reason, it is desirable to make the value of the reactance arms as high as possible. However, it can also be seen from Fig. 11-17 that the transmitter *A* drives the antenna loads through these reactances, and for this reason it is desirable to keep this reactance as low as possible. Actually, both of these conditions can be satisfied by tuning the bridge reactance arms to parallel resonance at the operating frequency of transmitter *B*. Thus for transmitter *B* the shunting effect of the bridge reactive arms is negligible owing to the high impedance of the parallel resonant circuit. At the same time, the

reactance of the inductive arms of the bridge can be made low enough for satisfactory operation of the transmitter A.

The basic circuit of Fig. 11-17 is satisfactory if the equipment all uses balanced lines. Since this is not usual, both the balanced transmitter

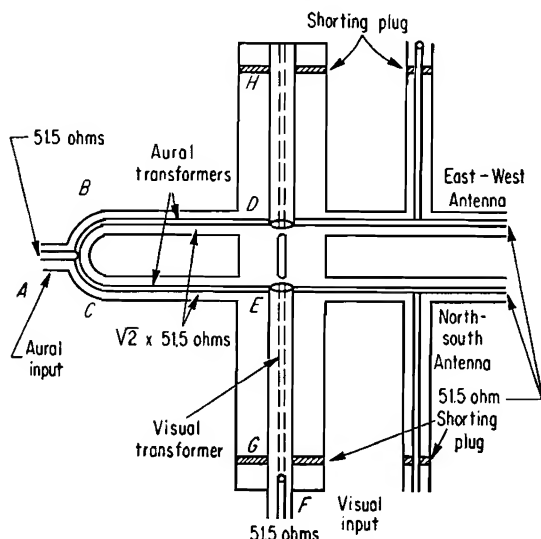


FIG. 11-18. A practical coaxial configuration of the balanced bridge diplexer.

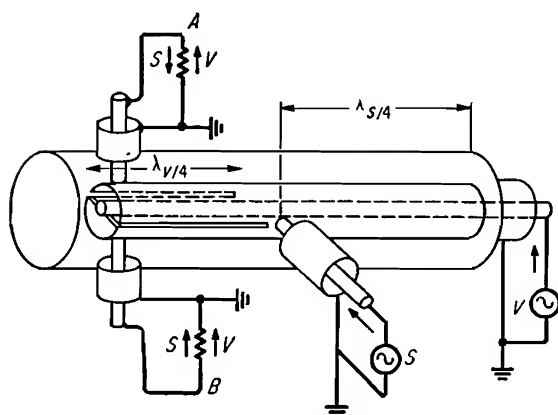


FIG. 11-19. A slotted bridge diplexer. (From *Wireless World*, March, 1950.)

inputs of Fig. 11-17 must be converted to single-ended or unbalanced inputs. A practical coaxial configuration of the balanced-bridge diplexer is shown in Fig. 11-18.

Slotted-bridge Diplexer. A second coaxial version of the bridge diplexer is shown in Fig. 11-19. While this configuration is more difficult to associate and visualize as being a bridge, it is a more compact form than

the type shown in Fig. 11-18. The inner conductor is split for a quarter wavelength at the frequency of operation. This section forms a split balun with one side connected to the inner conductor of the input transmission line. Thus the single-ended visual transmitter input is converted to a push-pull input as required by the equivalent bridge circuit. Since the aural transmitter input is made at the bottom of the split inner line section, the aural energy is fed to the output line in parallel as required by the equivalent bridge circuit.

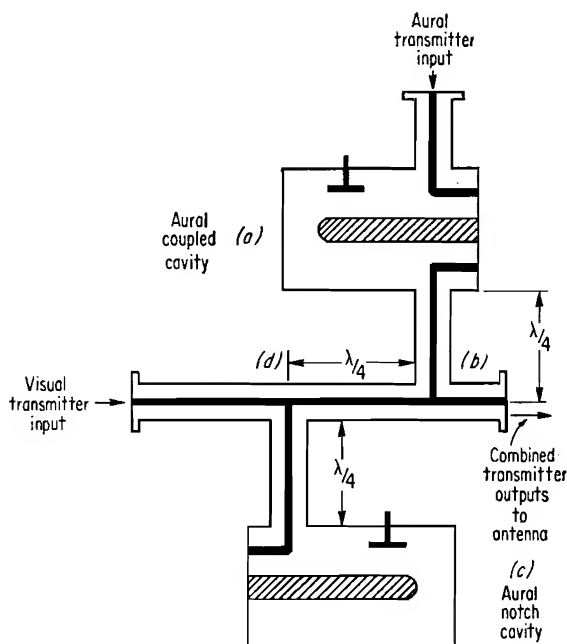


FIG. 11-20. The notch diplexer.

The Notch Diplexer. The notch diplexer provides a single-output transmission line for the two transmitter inputs. This can provide substantial savings (when compared to the bridge diplexer) in the cost of the transmission line between the diplexer and a single-feed antenna.

A typical notch diplexer is shown in schematic form in Fig. 11-20. The operation of this type of diplexer is best understood by considering the action of a signal supplied to each of the inputs. It must be first assumed that each of the cavities is tuned to the aural input frequency. Now, when a signal is impressed on the aural input line of Fig. 11-20, the signal will pass through the aural cavity A and appear at the output at point B. At point B the signal can potentially divide between the path leading to the load and the path leading back into the visual transmitter

input. However, since the aural cavity at C is tuned to the aural frequency, a short circuit appears across the transmission line at point D due to the quarter-wavelength transmission-line section between the aural cavity at C and the transmission line at D . Similarly, a high impedance to the aural frequency appears across the transmission line at B owing to the quarter-wavelength line between points D and B .

Now consider a signal at the visual input to the diplexer. Since the aural cavity at C is not tuned to the visual frequency, it presents a low impedance to a visual input frequency. However, because of the quarter-wavelength section between points D and C , the aural cavity merely acts like a high impedance across the transmission line. The cavity at A performs in a similar way. The cavity at A also prevents the visual signal from appearing at the aural input.

11-8. The Magic T. The propagation of electromagnetic waves in a waveguide can be used to form an effective diplexer. One configuration, shown in Fig. 11-21, is called the *magic T*. The magic T provides a dual output for the load.

The operation of the magic T can be easily visualized by separating it into its component parts. The magic T is actually composed of two types of waveguide junctions. One of these junctions together with the lines of the electric field within the waveguide is shown in Fig. 11-22. This type is known as the *series junction*. The second junction making up the magic T is the *shunt junction*. This type, with the resulting electric and magnetic field is shown in Fig. 11-23. If these two junctions are combined, the magic T shown in Fig. 11-21 will be formed. From the fields shown in Fig. 11-22 and Fig. 11-23, it can be seen that the field from one junction will not propagate up the waveguide forming the opposite type of junction and vice versa. In this way, the inputs to the waveguide shown in Fig. 11-21 are isolated; yet their outputs are combined.

Theoretically this type of diplexer is basically insensitive to frequency (within the limitations of the waveguide). However, when a junction of either type described is made, an input mismatch will result. This mismatch is eliminated by an adjustable post shown in Fig. 11-21. In practice, these matching methods make the magic T diplexer somewhat frequency-sensitive.

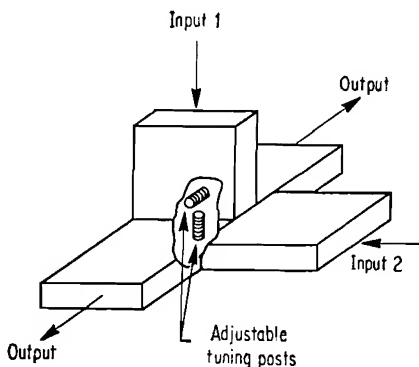


FIG. 11-21. The magic T.

11-9. The Coaxial Hybrid Ring. The coaxial hybrid ring, which is also known as the "rat race," provides an inherently unbalanced input and output connection. This makes its use convenient with coaxial

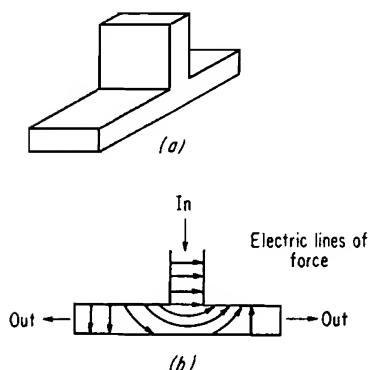


FIG. 11-22. The development of the magic T. The series waveguide junction.

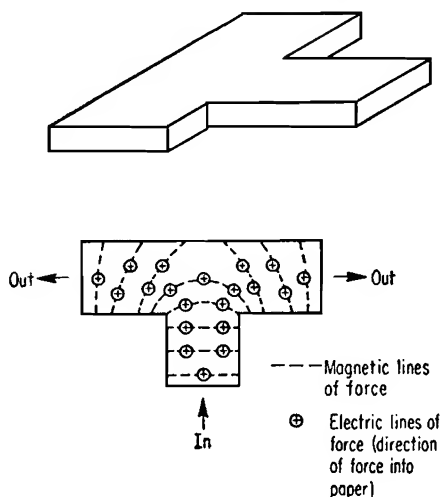


FIG. 11-23. The development of the magic T. The shunt waveguide junction.

transmission line. This is in contrast to the bridge circuits previously described which basically required a balanced input connection and which required the use of a balun to convert the balanced input to an unbalanced connection for use with coaxial cable.

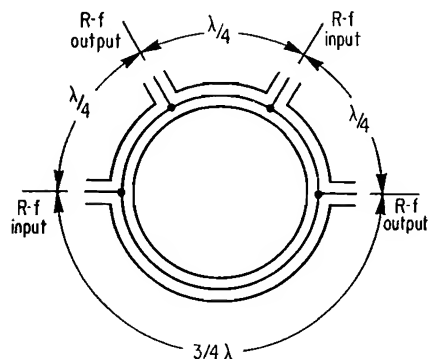


FIG. 11-24. Schematic of typical coaxial hybrid ring.

combinations of spacing between the various input and output connections can be derived to accomplish the same diplexing action.

11-10. Common Frequency Diplexing of Transmitters. When transmitters of the same frequency are to be combined into one output, some

To prevent input r-f mismatches, the coaxial lines making up the hybrid ring have a characteristic impedance equal to the square root of two times the impedance of one of the arms.

The operation of the hybrid ring can be easily deduced from the schematic of Fig. 11-24, which illustrates the typical wavelength spacings between the various input and output connections. The com-

additional restrictions are imposed on the type and characteristics of the diplexer. For example, with both inputs to the diplexer on the same frequency, it is impossible to use the notch-type diplexer since no isolation between the transmitters would be provided. In addition, most combining schemes require that the signals to be combined must be accurately controlled in phase and amplitude.

The most popular method of combining two identical frequency transmitters is based on the bridge. The principle of such a bridge is shown in Fig. 11-25. This circuit is similar to that previously shown for the bridge diplexer in Fig. 11-17. However, in this case, one of the resistive arms for the load output is replaced by a resistive dummy load. Under properly adjusted conditions, the dummy resistor does not dissipate any power. When current from transmitter A is adjusted to be exactly equal in amplitude and opposite in phase to the current from transmitter B, the currents from the two input transmitters A and B will

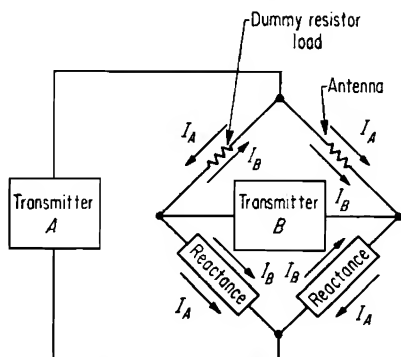


FIG. 11-25. Basic bridge method of diplexing two transmitters operating on the same frequency. I_A is the r-f current from transmitter A. I_B is the r-f current from transmitter B.

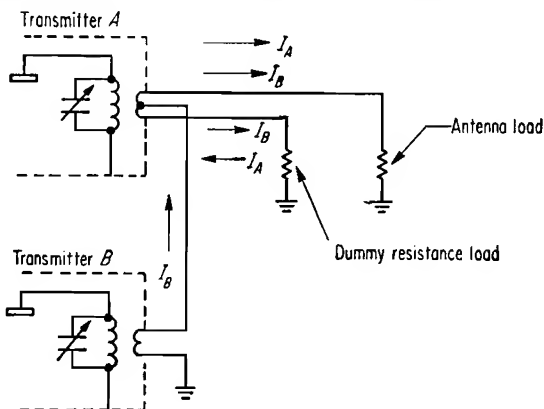


FIG. 11-26. Practical arrangement of diplexing at low and medium frequencies. I_A is r-f current from transmitter A. I_B is r-f current from transmitter B.

cancel; so the power dissipated in the dummy resistors will be zero. Similarly, the currents from transmitters A and B will add through the actual load resistance.

The basic bridge shown in Fig. 11-25 can be used directly at the lower frequencies by the practical circuit shown in Fig. 11-26. However, this

bridge arrangement becomes a little awkward in the vhf region since one of the inputs is unbalanced and it requires transmitter connections to widely spaced points in the bridge. In practice, the coaxial bridge using a hybrid-ring arrangement, such as is shown in Fig. 11-27, is commonly used. This permits the use of the more standard coaxial transmission line for each of the transmitters. The purpose of making one of the arms of the bridge three-quarters of a wavelength ($3\lambda/4$) is to permit the phase cancellation of the two transmitter signals at the dummy load resistor. If the arms of the bridge were all made one-

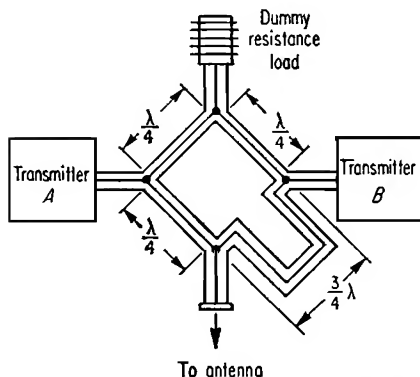


FIG. 11-27. A coaxial bridge using a hybrid ring for diplexing.

quarter wavelength, the currents through the dummy load resistor would equal those through the antenna load. Since a half-wavelength section of transmission line acts like a one-to-one transformer with a 180° phase shift, including such a section in one arm of the bridge permits the two transmitter signals to be canceled at the dummy load. By increasing the arm an additional quarter-wavelength to a total of three-quarters wavelength, an amplitude balance of the bridge is also achieved. Ideally

the power dissipated by the dummy load resistor under these conditions would be zero. However, in the event one of the input transmitters should fail, the dummy resistor is normally rated to be capable of dissipating one-half the power of one of the transmitters.

Diplexers are used to combine groups of transmitters to obtain a single combined power output. Arrangements similar to that shown in Fig. 11-28 have been used. The number of transmitters must be an integral power of 2, i.e., 2, 4, 8, 16,

11-11. Special Adjustment of Combined Transmitters. Two adjustments are required for maximum efficiency from a diplexer used for combining transmitter power outputs. The first requirement is that the two transmitter power outputs be equal. This is generally achieved through the use of the output coupling control, the input coupling control, or by the control of the d-c input power.

The second requirement is that the phase of the two signals should be opposite at the diplexer dummy load. This requires a phase control at one of the transmitters. In the vhf and uhf regions, a line stretcher is conveniently used for this purpose. This device is merely a coaxial line section that is adjustable in length. A cross section of a typical

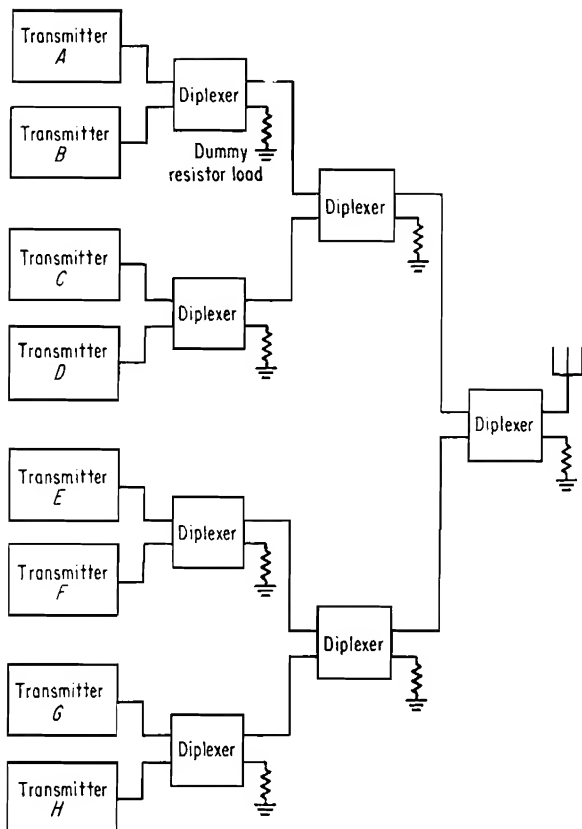


FIG. 11-28. A method of diplexing groups of transmitters.

line stretcher is shown in Fig. 11-29. This phase adjustment is relatively uncritical. An error of 30° in this adjustment produces an approximate decrease in the power output of only 3 per cent.

11-12. Power Dividers. As its name implies, the power divider is a device that will divide the power between two loads. Typically, the power divider is an adjustable device which makes it convenient for use as an r-f power control. For example, it is often required to control the r-f drive power into an active r-f load such as the input cavity of a klystron amplifier. At the same time, it may be necessary to provide a constant load to the

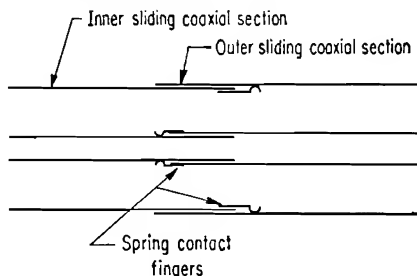
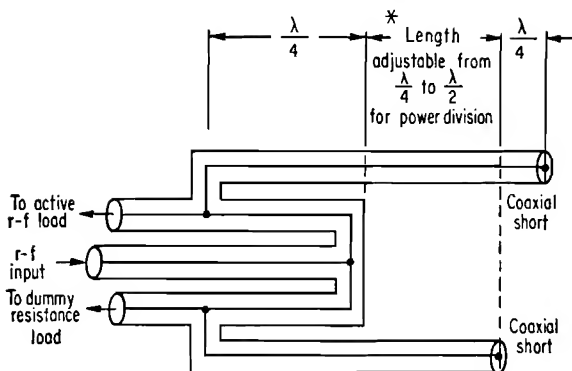


FIG. 11-29. A basic construction of a coaxial-line stretcher.



* When this length is adjusted to $\lambda/2$, all r-f input power will appear at active r-f load.
 When this length is adjusted to $\lambda/4$, all r-f input power will appear at the dummy load.

FIG. 11-30. A coaxial power divider.

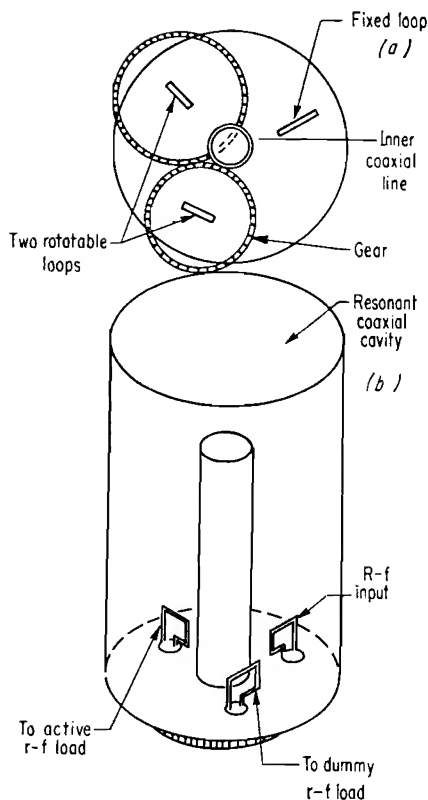


FIG. 11-31. A resonant-cavity power divider. (a) Cross section; (b) cutaway view.

driver—this in order to maintain a constant driver bandwidth or simply to prevent the driver output tube from exceeding its ratings owing to a varying load. Two typical power dividers which will accomplish both of these purposes are shown in Fig. 11-30 and Fig. 11-31.

The coaxial power divider shown in Fig. 11-30 will proportionately divide the power between the active and dummy loads according to the position of the adjustable transmission-line section shown in the figure. As this length is varied from one-quarter to one-half wavelength, the power will divide between the dummy and active loads. This type of power divider requires that the fixed transmission-line lengths indicated be previously adjusted to the proper lengths according to the expected operating frequency. These sections are usually made adjustable and may be set with the aid of a frequency-vs.-length chart.

The second type of power divider shown in Fig. 11-31 uses the fixed 90° relationship between two loops which are mechanically ganged together by gears and are rotatable to obtain the power division. Thus, as the gang loops are rotated, one loop will gradually increase its coupling

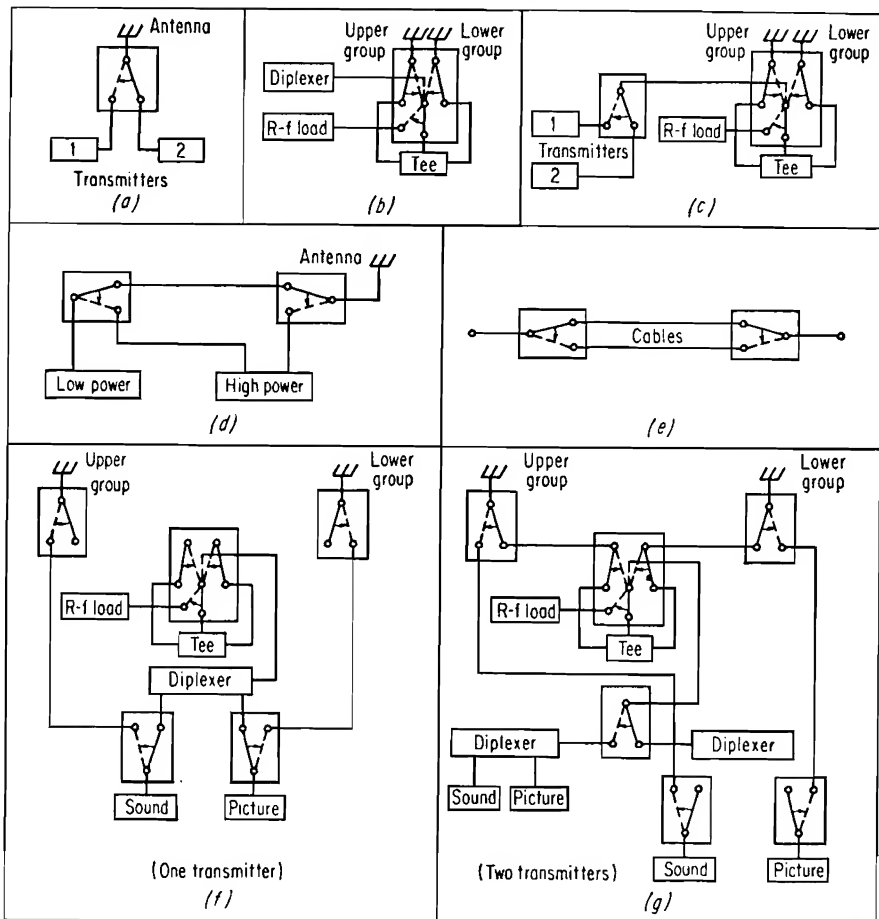


FIG. 11-32. Typical applications for coaxial switches. (a) Transmitter switching; (b) antenna sectionalizing dummy load switching; (c) simple switching combination; (d) power cutback; (e) spare cable switching; (f) typical one-transmitter switching arrangement; (g) typical two-transmitter switching arrangement.

to the r-f cavity as the other loop gradually decreases its coupling. Energy is introduced into the cavity by means of a fixed loop. The resonant frequency of the coaxial cavity must be adjusted to correspond to the operating frequency if proper operation and good input match is to be obtained.

11-13. Transmission-line Switching. The need often arises to switch or transfer the output of the transmitter from one antenna to another, or

to transfer the transmitter output to a dummy load for test purposes. Conversely, it is often desired to transfer the antenna from the main transmitter to a stand-by transmitter in the event of a failure.

Since valuable transmitter time can be lost in making a manual transfer of the transmission line, it is convenient to make use of a manual transmission-line transfer panel or an electrically operated transmission-line switch. Some transmission-line switching arrangements are illustrated in Fig. 11-32.

Manual Coaxial Transfer Panel. A manual transfer panel is made up of a U-shaped link constructed of two right-angle elbows, which may be connected to a transmission line as shown in Fig. 11-33. The outer

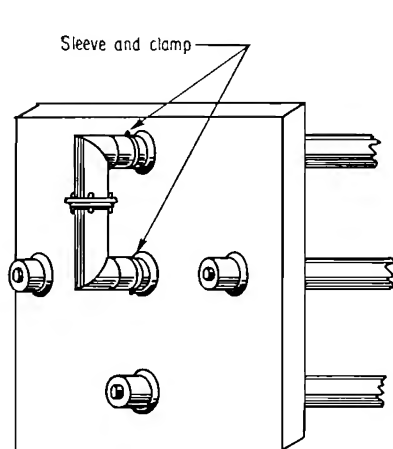


FIG. 11-33. A typical coaxial transmission-line manual transfer panel.

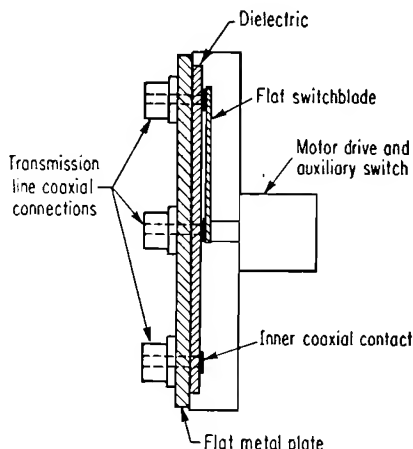


FIG. 11-34. A typical high-power coaxial transmission line switch.

coaxial connection is made with a sleeve clamped to the outside to facilitate rapid removal and transfer when desired.

Electrically Operated Coaxial Switches. The electrically operated transmission-line switch performs in much the same manner as a simple single-pole tap switch, except that in this case the circuit through the switch must have a characteristic impedance equal to that of the transmission line. This is done by considering the flat blade of the switch as a flat transmission line with the case as ground. A switch utilizing this construction is shown in Fig. 11-34.

One method of ensuring intimate contact between the flat switch-blade and the contact button is to make the blade of silver-plated steel and to make the contact a permanent magnet. When the switch arm is positioned over the contact by the motor, the magnet will assure good contact.

The switch motor must also operate an auxiliary switch, not only for internal control of the operating motor, but also for the transmitter interlocking functions. Normally, the auxiliary switch is interlocked with the transmitter to shut off the transmitter power while the r-f switching is taking place.

11-14. Microwave Ferrites. A large group of r-f components for the microwave frequencies depend on the characteristics of ferrites for their operation. Ferrite is a material which has the magnetic permeability of magnetic materials but a resistivity 10^{16} times that of iron. This permits the use of ferrites up to the microwave frequency with only a negligible eddy-current loss.

Chemically a ferrite is a compound of iron, oxygen, and one other metal. Its chemical formula is MFe_2O_4 , where M is the other metal. This other metal may be either cadmium, cobalt, copper, magnesium, or nickel. A microwave ferrite typically consists of a solid solution of two or more ferrites of different composition.

Ferrite applications may be divided into the reciprocal and non-reciprocal classes. In the reciprocal class, the phase or amplitude of a microwave signal is made dependent on an externally applied magnetic field. In the nonreciprocal class, the phase or amplitude of the signal depends on the direction of the signals through the device.

The nonreciprocal properties of ferrites are due to ferromagnetic resonance absorption and Faraday rotation. Briefly, ferromagnetic resonance absorption results from the application of a frequency to the ferrite which equals the ferromagnetic resonance frequency. When this occurs, the frequency applied to the ferrite is absorbed as heat by the ferrite material.

Faraday rotation is the rotation of a wave passing through a ferrite in a direction parallel to an applied steady magnetic field. If this wave is reflected back through the material, it will be again rotated in the same direction as the incident wave. Thus the total rotation will be twice the original. This makes the ferrite nonreciprocal in action.

The two classes of ferrites (reciprocal and nonreciprocal) find use as isolators, circulators, r-f switches, and amplitude, frequency, and ssb modulators.

11-15. Isolators. The isolator is a one-way transmission device which makes the use of the nonreciprocal characteristics of ferrite. Its primary use is to isolate an r-f generator from its load. This isolation can be accomplished by a resistive pad; however, this also results in a loss in power in the forward power direction. The ferrite isolator has a negligible power loss in the forward power direction and up to 30 db loss in the reverse power direction.

A typical application of the isolator is at the output of a klystron

oscillator feeding a long waveguide. If the oscillator frequency is varied by some intentional means, the reactance seen by the oscillator will also vary due to any slight mismatch in the transmission line. This effect, known as the *long-lines effect*, will cause the oscillator frequency to be "pulled" as a result of the varying reactive load. This application is illustrated in Fig. 11-35.

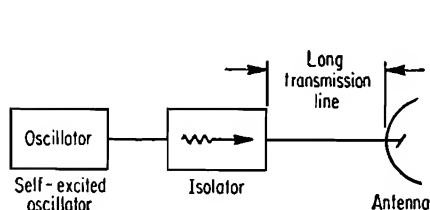


FIG. 11-35. A typical application of an isolator in a long transmission line.

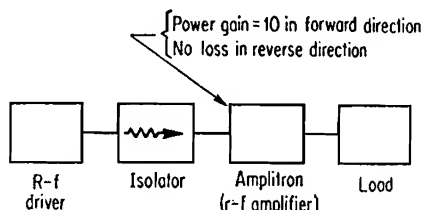


FIG. 11-36. An isolator used with an amplatron r-f amplifier.

Another typical application for the isolator is between the driver and the input of an amplatron r-f amplifier. An isolator is required in this application since any signal reflected from the load as a result of mismatching will pass through the amplatron undiminished. Without an isolator, this power would appear reflected back into the driver. This condition is illustrated in Fig. 11-36.

A typical isolator using the principle of Faraday rotation is illustrated in Fig. 11-37. In high-power applications, some consideration is given

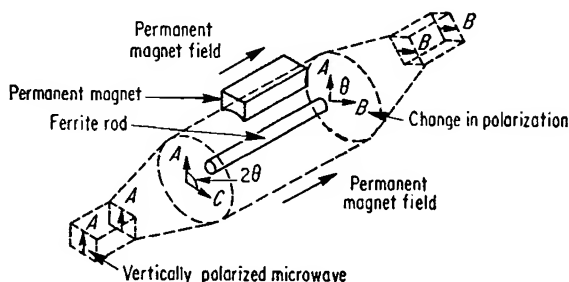


FIG. 11-37. A ferrite isolator using Faraday rotation. Signal A passes through ferrite rod, is rotated θ° , and emerges as signal B. If B is reflected it will be rotated another θ° for a total of 2θ . (Courtesy of Lenkurt Corporation.)

to cooling the ferrite by convection, forced-air, or liquid cooling. The use of flat slabs of ferrite as shown in the isolator of Fig. 11-38 simplifies the cooling of the ferrite.

11-16. Circulators. Circulators are directional r-f devices that will pass the r-f power from one terminal of the device to another, in sequence, until the power is absorbed at one of the terminals. For example, in Fig. 11-39 showing the schematic symbol of a four-port circulator, a

signal entering at port 1 will emerge at port 2. Similarly, a signal entering at port 2, will emerge from the circulator at port 3, and so on for all the ports. This permits the use of a circulator as a duplexer for connecting a transmitter and receiver to a common antenna. This is

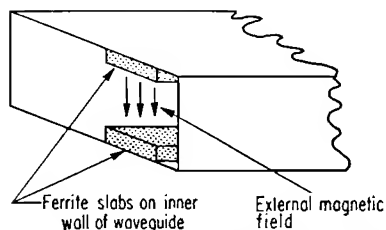


FIG. 11-38. Typical high-power ferrite isolator.

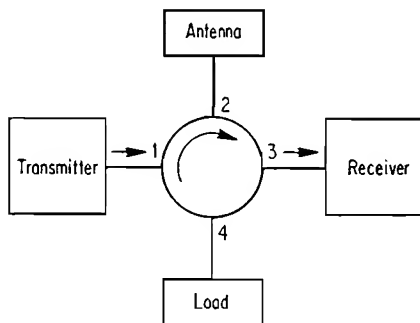


FIG. 11-39. A circulator used as a duplexer.

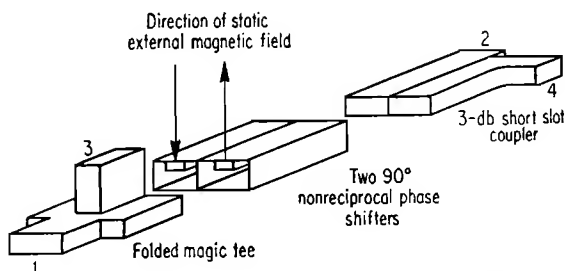


FIG. 11-40. A typical high-power circulator. (Courtesy of Microwave Journal.)

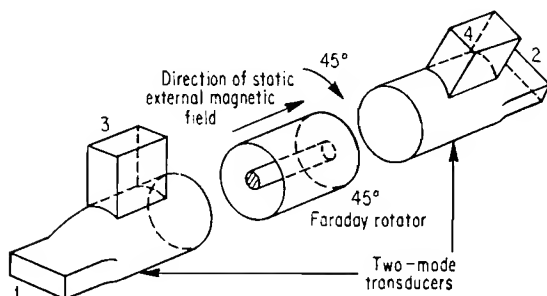


FIG. 11-41. A four-port Faraday rotation circulator. (Courtesy of Microwave Journal.)

illustrated in Fig. 11-39. This system of duplexing offers the advantage of isolating the transmitter from any effects of antenna mismatch. In addition, any r-f, fed into the transmission line from the local oscillator of the receiver, will not pass out to the antenna but will be dissipated in the load. This system also permits duplexing of c-w (continuous-wave)

signals. Figure 11-40 shows the essentials of the construction of a typical high-power circulator, while Fig. 11-41 shows the typical construction of a circulator used for medium- and low-power applications.

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TRANSMITTER CHARACTERISTICS

12-1. Transmitter Applications. A complete transmitter will be an assembly of components described in earlier chapters. The characteristics of the transmitter and in turn the components will be determined by the service requirements. In many cases the nature of the transmitted signal such as the radiated power, frequency tolerance, bandwidth occupied, and other qualities will be limited by government regulation.

Transmitter applications may be broadly divided into the following categories:

1. Navigation
2. Radar
3. Ground communication
4. Mobile communication
5. Broadcast

A discussion of the requirements for each category would lead to system engineering considerations which are beyond the scope of this book. However, a few features and special problems associated with transmitters for the various classes of service will be discussed. The list of references will lead to more detailed discussions.

12-2. Navigation. Transmitters which are used to assist in the navigation of a ship or an aircraft may be located on the ground, as with beacons, ranges, and instrument landing systems, or may be located on the vehicle as is the case of the interrogating transmitter in distance-measuring systems. There are a number of sources of information on electronic navigation systems (1-7). Electronic systems that are used specifically for the direction of aircraft have been called *avigation* systems.

In many cases the transmitters are simple c-w or a-m units, and the special characteristics which make the system suitable for navigation purposes are achieved entirely in the antenna system. The A-N low-frequency range (8) and the v-h-f omnirange (VOR)—see Ref. 9—are typical examples. In the case of the *localizer* and *glide-path* transmitters which are part of the instrument-landing system (ILS), modulation of

the signal is achieved in the antenna system by mechanical means and the transmitters are simple c-w units. Reliability is of the utmost importance in this class of equipment and the transmitters are kept as simple as possible.

Loran. Loran is a long-range navigation system in which the difference in time of arrival of pulses from widely spaced transmitters is measured at the receiver in the ship or aircraft (10,11). The accurate timing of the pulses generated by each transmitter is accomplished in the terminal, and the transmitters themselves are quite simple (12). The transmitter stability need not be better than 0.1 per cent to ensure that the signal falls within the receiver pass band which is 75 kc wide, and early loran transmitters were not crystal-controlled. In later systems the relative phase of r-f cycles within the pulse is measured and crystal control became necessary.

It is important that the rise time of the pulses in a loran transmitter is not too fast; otherwise excessive sideband radiation occurs. The rise time must be at least as long as 20 μ sec. Standard loran operates in the 1,750- to 1,950-kc band. An improved system called *Cytac* operates in the region of 100 kc (13).

A British position-finding system called *Decca* (14) operates in the vicinity of 100 kc and involves the phase comparison of c-w signals which are on different frequencies but related to a common higher frequency. The three slave transmitters are simple c-w units.

Tacan and DMET. The tacan (tactical air navigation) and the DMET (distance measuring element of Tacan—see Ref. 15) systems employ air-borne interrogating transmitters and responding transmitters on the ground (16,17). These are pulse transmitters which send pulses having a duration of 3.5 μ sec. The ground-beacon transmitter employs a *constant-duty-cycle* type of modulation, and about 2,700 pulses per sec are sent, whether the beacon is being interrogated or not. This permits the beacon to be used for bearing as well as distance measurement. The ground beacon operates on a fixed frequency and has a power output (peak) between 5 and 30 kw. The air-borne transmitter has a peak output of 1.5 kw and must be immediately resettable to any one of 126 channels.

12-3. Radar. The characteristics of a radar transmitter depend on the application, and there are many variations (18). Frequencies commonly used for radar are given letter designations which are given in Table 12-1. Nomographs are available for the quick determination of the transmitter power required for a desired range and frequency (19).

In general, as the desired range increases, the r-f frequency comes down, the pulse-repetition rate comes down and both the pulse width and peak power increase. As the desired *resolution* or the ability to separate small

TABLE 12-1. BAND CODE LETTERS VS. FREQUENCY*

P Band			S Band			C Band (designa- tion includes Sz through Xr from 3.9-6.2 kMc/sec			K Band			Q Band				
Fre- quency	λ		Sub	Fre- quency	λ	X Band			Sub	Fre- quency	λ	Sub	Fre- quency	λ		
0.225	133.3		E	1.55	19.3	Sub	Fre- quency	λ	P	10.90	2.75	A	36.00	0.834		
0.390	76.9			1.65	18.2				S	12.25	2.45		38.00	0.790		
Uhf television band Channels 14 through 83, 0.470- 0.890 kMc/sec			F	1.65	18.2					12.25	2.45	B	38.00	0.790		
				1.85	16.2					13.25	2.26		40.00	0.750		
			T	1.85	16.2	E	13.25	2.26	C	40.00	0.750					
				2.00	15.0		14.25	2.10		42.00	0.715					
L Band			C	2.00	15.0	A	5.20	5.77	C	14.25	2.10	D	42.00	0.715		
				2.40	12.5		5.50	5.45		15.35	1.95		44.00	0.682		
			Q	2.40	12.5	Q	5.50	5.45	U	15.35	1.95	E	44.00	0.682		
				2.60	11.5		5.75	5.22		17.25	1.74		46.00	0.652		
			Y	2.60	11.5	Y	5.75	5.22	T	17.25	1.74	V Band				
				2.70	11.1		6.20	4.84		20.50	1.46					
			G	2.70	11.1	D	6.20	4.84	Q	20.50	1.46	Sub	Fre- quency	λ		
				2.90	10.3		6.25	4.80		24.50	1.22					
			S	2.90	10.3	B	6.25	4.80	R	24.50	1.22					
				3.10	9.68		6.90	4.35		26.50	1.13					
			A	3.10	9.68	R	6.90	4.35	M	26.50	1.13	A	46.00	0.652		
				3.40	8.83		7.00	4.29		28.50	1.05		48.00	0.625		
						C	7.00	4.29	N	28.50	1.05	B	48.00	0.625		
			W	3.40	8.83		8.50	3.53		30.70	0.977		50.00	0.600		
				3.70	8.11	L	8.50	3.53	L	30.70	0.977	C	50.00	0.600		
			H	3.70	8.11		9.00	3.33		33.00	0.909	D	52.00	0.577		
				3.90	7.69	S	9.00	3.33	A	33.00	0.909	E	54.00	0.556		
			Z	3.90	7.69		9.60	3.13		36.00	0.834		56.00	0.536		
				4.20	7.15	X	9.60	3.13	K1 Band designa- tion includes Kv through Kq 15.35- 24.5 kMc/sec .350-.530 kMc also called J band							
			D	4.20	7.15		10.00	3.00								
				5.20	5.77	F	10.00	3.00								
							10.25	2.93								
						K	10.25	2.93								
							10.90	2.75								
X	0.950	31.6														
	1.150	26.1														
K	1.150	26.1														
	1.350	22.2														
F	1.350	22.2														
	1.450	20.7														
Z	1.450	20.7														
	1.550	19.3														

Legend: Sub = identifying subletter.

Frequency = band limits, kMc per sec

 λ = equivalent wavelength, cm

* Courtesy Vectron Products Division of the Itek Corp.

objects increases, the pulse width must be reduced, and the pulse-repetition rate is increased. Transmitter characteristics for a few representative types of radar are given in Table 12-2.

TABLE 12-2. TRANSMITTER CHARACTERISTICS FOR SOME TYPES OF RADAR

Type	Frequency, Mc	PRF, pps	Pulse width, μ sec	Duty ratio	Peak power, kw
Shipboard search.....	220	60	5	0.0003	200
Microwave early warning.....	2,800	400	2	0.0008	1,000
Mobile land-based search.....	3,000	1,700	0.8	0.0014	300
Aircraft height finder.....	3,000	1,000	1	0.001	60
Air-borne weather.....	5,400	400	2	0.0008	75
Shipboard navigation.....	9,400	800	0.5	0.0004	15
Precision aircraft approach.....	10,000	2,000	0.5	0.001	20
Airport surface detection.....	24,000	14,400	0.02	0.0003	50

The receiver associated with a radar transmitter must be activated between pulses and must not be damaged when the transmitter comes on. The same antenna is generally used for transmitting and receiving, and the transfer or simultaneous connection without damage to the receiver is achieved with a *duplexer* (20). The distinction between a *duplexer* and a *diplexer* is that the useful energy passes in opposite directions in a *duplexer* and in the same direction in a *diplexer*. A common arrangement used in radar transmitters is the use of a transmit-receive (TR) and an anti-TR (ATR) switching tube. The tubes are enclosed spark gaps which "fire" when the transmitter is generating r-f energy. When the TR tube fires, it short-circuits the line to the receiver. When the ATR tube is not fired, the open circuit is transformed to appear as a short circuit at the transmitter output.

The Doppler effect is used in certain types of radar, and in this case very long pulses or continuous waves may be used (21,22). C-w transmitters are used in automobile speed-checking radars (23). Frequency modulation is sometimes used in Doppler radars and is also used in low-altitude radio altimeters (24).

12-4. Telegraph Transmitters. Telegraph transmitters are used at *fixed* point-to-point stations and at *coastal* stations which communicate with ships at sea.

The two basic types of emission on telegraphy are on-off keying (A1) and frequency-shift keying (F1) (25). Hand-keyed International Morse is used for communication between coast stations and ships. It is desirable for each operator to be able to listen through his own sending

and be ready to stop when the other operator *breaks* him. For this reason on-off keying is used in this type of service.

Frequency-shift keying is extensively used in *record* communications. The usual recording machine is a teleprinter, and the Baudot seven-unit code is commonly used.

In the simplest fsk systems, the transmitter frequency is shifted between two values, usually separated by 850 cps. The *mark* (printer idling) signal is 425 cps above the assigned carried frequency. The principal considerations in the selection of a method of frequency-shift keying are the frequency stability of the mark and space signals and the

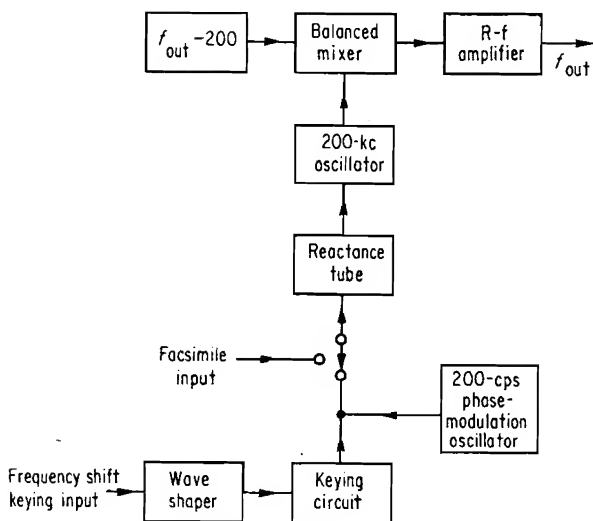


FIG. 12-1. Reactance-tube frequency-shift keyer.

amplitude of the spurious outputs. Excessive sidebands will be generated if the mark-space transition is too rapid (26,27).

Pulled-crystal Method. The frequency of a crystal oscillator may be changed by keying a capacitor across the crystal or switching crystals; but transients will be generated. If a keying tube is used to change the shunt capacity across the crystal, waveshaping circuits can be used to increase the mark-space transition time.

Low-frequency Oscillator-Mixer. Another method is to shift the frequency of a low-frequency self-excited oscillator with a reactance-tube circuit. The oscillator output (100 to 200-kc range) is mixed in a balanced mixer with a stable source which is offset from the desired output frequency by an amount equal to the keyed oscillator frequency. Waveshaping circuits ahead of the reactance tube can be designed to achieve any desired type of mark-space transition. This type of circuit can also be used for facsimile transmission where a smooth and continu-

ous variation from black to white is required. The block diagram of such a keyer is shown in Fig. 12-1.

Phase-locked Oscillator. A phase-locked oscillator can be frequency-shift-keyed by keying the reference oscillator. Shaping of the mark-space transition is accomplished in the loop filter, as shown in Fig. 12-2.

Error-reducing Methods. One method of reducing telegraph transmission errors is the automatic-error-correcting (ARQ) system which was described by Van Duuren (28). A special seven-unit code is used in

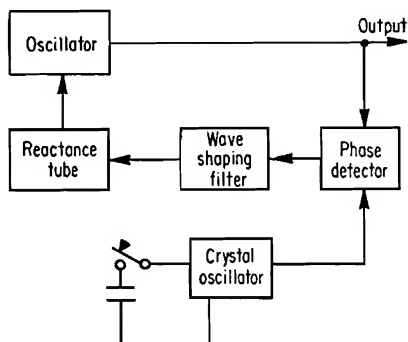


FIG. 12-2. Frequency-shift keying of phase-locked oscillator.

which each character has three marking and four spacing elements. If the received signal does not have this ratio, a signal is sent back to stop the transmitter and repeat the character.

The other approach is to devise keying techniques which give improved performance in the presence of noise. A number of techniques have been proposed for this purpose.

The *predicted-wave* Teletype¹ method (29) gives a few decibels improvement over 850-cps frequency-shift

keying and uses special detection and filtering techniques at the receiver. At the transmitter a synchronizing signal which is half-way between the mark and space frequencies is sent once for each character.

Phase-shift Teletype. This form of transmission gives about 10 db improvement over 850-cps frequency-shift-keying (30). Here the steady-state frequency is the same for both mark and space, but the phase of the transmitted signal shifts 180° for a mark-space transition. There is, of course, a frequency change during the phase change, but the required transmission bandwidth can be minimized if the phase is not changed too rapidly. With both the predicted-wave and phase-shift Teletype transmission systems, it is necessary to use a *synchronous* code which means that all the code elements have equal length. In the standard Baudot seven-unit code the "stop" pulse is 142 per cent longer than each of the other pulses. 180° phase keying is called *biphase* keying.

Single-sideband Techniques. A number of telegraph channels can be transmitted simultaneously by modulating the transmitter with a number of subcarriers, each of which is frequency-modulated. Economy of transmitter power and also of frequency spectrum can be improved by using ssb transmission of the subcarriers (31,32). Keying information from message-handling centers is often sent to transmitting stations by means of keyed tones. These are usually rectified to obtain direct

¹ Teletype Corp. of America trade name.

current for keying the transmitters. One simple method of converting to ssb subcarrier operation is to eliminate rectification of the tones and to feed them directly into an ssb transmitter. The audio subcarriers in general use for this service are spaced 170 cps apart between 425 and 3,315 cps for keying speeds up to 100 words per min. For higher speeds, 340-cps spacing is used. Each subcarrier is shifted 30 to 42.5 cps for the 170-cps spaced systems and 50 to 170 cps for the 340-cps spaced systems.

Frequency-stepping FSK. Multipath distortion is a serious problem on high-frequency circuits. One method of dealing with this problem has been to "dodge" the multipath signal by shifting the transmitter to a new average frequency for each code element (33). It has been found that it is enough to use three steps before returning to the first frequency again. Maximum multipath delays are about 12 msec. The receiver, must of course, be stepped simultaneously to avoid reception of the multipath interference signal. Spacings of 300 cps for the shift and 600 cps for the steps have been used.

Multiplex Keyers. The transmission of two or more information channels on a common carrier is called *multiplexing*. The use of a number of subcarriers is called the *frequency-division* method and has been discussed briefly. In the *time-division* method the different messages are sent in successive time slots, and some form of electrical or mechanical *decommutator* is required at the receiver. A complete discussion of multiplexing methods is beyond the scope of this book. A number of four-channel time-division equipments are in use for sending teletype information on fsk transmitters (34).

If an fsk transmitter is arranged to send on one of four different frequencies at a time, two independent channels using different codes at different speeds can be sent (35). Table 12-3 shows the combinations.

TABLE 12-3. CHANNEL COMBINATIONS POSSIBLE WITH FOUR DIFFERENT FSK FREQUENCIES

Frequency	Channel A	Channel B
f_1	Mark	Mark
f_2	Mark	Space
f_3	Space	Mark
f_4	Space	Space

Shifting from f_1 to f_2 changes channel B from mark to space but does not affect channel A.

In the Kineplex¹ system, a number of subcarriers are used, and two channels of information are sent on each subcarrier by causing the sub-

¹ Collins Radio Company trade name.

carrier to assume one of four conditions of phase relative to the preceding code element (36). Phase-shift keying employing four possible conditions is called *quadriphase* keying.

Keying of VLF Transmitters. A special problem arises in the keying of VLF transmitters which operate in the frequency range from 15 to 35 kc at output powers from 250 to 1,000 kw. The difficulty is that practical antennas have resistances of less than 1 ohm and reactances of 100 ohms or more. If on-off keying is used, it takes up to 40 msec to charge the antenna up to 95 per cent of the full power value. This limits keying speeds to about 18 words per min. Frequency-shift keying also presents a problem because the half-power bandwidth of a typical antenna is less than 30 cps. It is necessary to retune the antenna when the frequency is shifted; this is accomplished by changing the inductance with a saturable reactor or changing the capacity with an electronic switch (37).

12-5. Telephone Transmitters. For many applications such as ground-to-aircraft communications in the h-f and vhf bands, conventional dsb modulation is used. There is, however, a continuing tendency to switch to single sideband (38). For transoceanic service in the h-f band typical transmitters use *independent sideband* (isb) with two phone channels on the upper sideband and two more on the lower sideband (39).

Land Stations. Land stations which communicate with mobile units in the 30- to 500-Mc range use narrow-band frequency modulation.

The signal-to-noise ratios at the receiving end are improved if the average modulation is kept as high as possible. In a typical speech waveform the peaks are about 15 db above the rms value; but it is found that the intelligibility is not seriously impaired if the peak-to-rms ratio is reduced by as much as 6 db. This is done by *clipping* the peaks. The clipping action usually generates new high-frequency components, and it is necessary to include a 3,000-cps low-pass filter after the clipper stage to avoid transmission of these components (40).

UHF and SHF Radio Links. Radio links are extensively used for point-to-point transmission of multichannel telephone and television. Line-of-sight links use up to 10 watts (41) and over-the-horizon transmitters may generate up to 50 kw, but levels between 1 and 10 kw are more commonly used. Frequency modulation is commonly used, but ssb (42) and pulse modulation have also been used to advantage (43). The transmitter output stages may use klystrons (44,45) or traveling-wave tubes (46,47).

Amateur Transmitters. A large number of communication transmitters are operated by radio amateurs. The equipment is adequately described in amateur handbooks (48,49).

12-6. Mobile Transmitters. A mobile transmitter is one which is installed on a vehicle or carried by a person.

Motor Vehicle. The frequency bands used by motor-vehicle transmitters has gradually moved upward as the demand has expanded. The 1,605- to 1,750-kc band was first used for radio dispatching. The next step was the 30- to 44-Mc band. After World War II there was a considerable shift of operations to the 152- to 174-Mc band, and more recently the 450- to 470-Mc band has been used. There has been a constant effort to reduce the bandwidth occupied by the transmitters and also the pass band of the receivers so that more stations can be handled without mutual interference. Transmitters authorized after November 1, 1958, and all transmitters after October 31, 1963, must conform to the limits shown in Table 12-4 (50).

TABLE 12-4. LIMITATIONS PLACED ON TRANSMITTERS

Frequency band, Mc	Bandwidth, kc	Maximum deviation	Frequency stability, %	
			Over 3 watts	Under 3 watts
25-50	20	5	0.002	0.005
50-150	40	15	0.0005	0.005
150-450	20	5	0.0005	0.005
450-1000	40	15	0.0005	0.005

Another requirement applicable after the above dates is that transmitters must be fitted with low-pass audio filters. The performance of the filter between 3 and 15 kc must be such that the attenuation referred to at 1 kc must be at least equal to

$$\text{Attenuation} = 40 \log \frac{f}{3} \quad \text{db}$$

where f = frequency in kilocycles. At frequencies above 15 kc the attenuation must be at least 28 db greater than that at 1 kc. Typical motor-vehicle transmitters have output powers between 10 and 50 watts.

The first extensive use of transistorized circuits in transmitters has been in mobile equipment where weight and power consumption are important (51). An important use of the transistor has been as a switching device to replace the mechanical vibrator (see Sec. 8-10).

Maritime Mobile. Transmitters for small boats are available for operation in the 1,600- to 4,500-kc range, but most operation is between 2,000 and 3,000 kc. Certain channels in the 152- to 174-Mc vhf band are also reserved for this service. Conventional amplitude modulation is used in the m-f transmitters.

Ocean-going ships are required to carry radiotelegraph transmitters which can send mcw (modulated continuous wave) or icw (interrupted continuous wave) on the 500-kc international distress frequency. Traffic is usually exchanged in the 415- to 490-kc range for short distances and on one of the maritime bands between 2 and 30 Mc for long distances.

Aircraft. Air-borne transmitters must have minimum size and weight and must be automatically tuned so that the pilot is not required to divert his attention to transmitter adjustment when changing to a new frequency.

For traffic control near airports and for short-distance enroute communication, the 118- to 132-Mc vhf band is used. For longer distances the 2- to 30-Mc range is used. Conventional amplitude modulation is used on both bands. Single sideband is also used in the h-f range. In high-speed aircraft, the Doppler effect becomes important at the higher frequencies. The shift in frequency due to Doppler is about 1 cps per Mc per Mach number. The receiver must be tuned to within 100 cps of the correct frequency to obtain 90 per cent intelligibility of speech for a signal-to-noise ratio of 16 db when ssb transmission is being used.

Typical aircraft transmitters vary in power level from 3 to 1,000 watts and operate from 27.5 volts d-c or 400 cps 115 volts a-c.

12-7. Telemetry Transmitters. Radio telemetry has important applications in the operation of unmanned aircraft, missiles, and earth satellites (52). Three radio telemetry bands in use in the United States are

216-260 Mc
1,435-1,535 Mc
2,200-2,300 Mc

Transmitters on earth satellites and outer-space probing vehicles have used narrow bands around 108 and 960 Mc (53).

A large quantity of data is required during the first few seconds of flight of a rocket, and consequently it was necessary to develop telemetry transmitters with a large information capacity. The principal systems in use are f-m/f-m (frequency division) and PDM/f-m or PCM/f-m (time division). In the f-m/f-m method, the transmitter carrier is modulated by a maximum deviation of ± 125 kc with up to 18 subcarriers which are between 400 and 70,000 cps (54). Time division may also be used with f-m/f-m by *commutating* one or more of the subcarriers to different sensing devices on the vehicle. Each subcarrier is frequency-modulated up to 7.5 per cent by the transducer which is sensing the information to be sent. In the PDM and PCM systems, the carrier is frequency-modulated by pulses, and information is contained in the width or coding of the pulses. A detailed discussion of these methods

would be considered as systems engineering and is beyond the scope of this work.

The actual transmitters are quite simple and are similar for all types of modulation. A deviation linearity of about 1 per cent over a range of ± 150 kc is required for the f-m/f-m applications. This is necessary to avoid intermodulation between the subcarriers. Linearity is less important with the time-division systems; about ± 35 kc deviation is used for PDM and ± 250 kc with PCM. Output powers of several watts have been used with telemetry transmitters, but, as the transmission is usually by line of sight, output powers as low as 10 mw can be used even at great distances, provided directional antennas and low-noise receivers are used on the ground (55,56).

12-8. Broadcast Transmitters. Broadcast equipment is characterized by a much closer control of performance by government regulation than other types (57). Power output must usually be maintained below low and high limits, and good frequency control is required. Modulation characteristics are also specified in detail.

A-M Transmitters. Amplitude modulation is used in the 535- to 1,605-kc band and also on certain channels in the 2- to 30-Mc range for international broadcasting. In Europe, Africa, and Asia the 150- to 285-kc band is also used for a-m broadcasting.

The maximum transmitter power-output level in the United States is 50 kw, but power outputs of up to 1 Mw (megawatt) are used in some countries (58).

Many broadcast transmitters are operated continuously; and at high-power levels the cost of electric power becomes important. Special high-efficiency linear amplifiers such as the Doherty amplifiers discussed in Sec. 6-8 are used in broadcast applications.

Compatible Single Sideband. The 10-kc channels in the standard broadcast band are butted against one another with no guard band between them. Adjacent channels are not normally allocated to the same area because of the *sideband monkey chatter* that would result from the desired carrier beating with sidebands from the next channel. More stations could be accommodated in the band if all transmitters were converted to single sideband; but this would involve modification of all receivers. A compatible system has been suggested by Kahn (59) in which one set of sidebands is attenuated about 30 db below the other. Conventional receivers may be used to receive the signal. The advantage is illustrated in Fig. 12-3.

F-M Transmitters. In the 88- to 108-Mc band the nominal peak deviation is ± 75 kc while the sound transmitters associated with television stations have a peak deviation of ± 25 kc.

The r-f amplifiers in f-m transmitters are operated as efficiently as

possible and are usually operated as class C amplifiers. In the uhf television band the aural transmitters may use klystrons.

Preemphasis of the higher audio frequencies is specified for f-m transmitters. The required amount for both f-m broadcast and TV sound

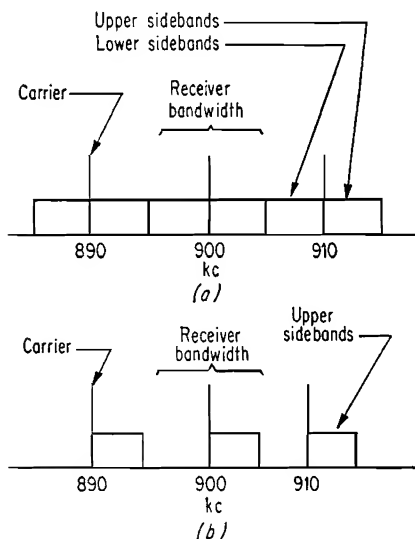


FIG. 12-3. Occupation of frequency spectrum in broadcast band. (a) Conventional double sideband; (b) compatible single sideband.

is that which will be obtained from an *RC* or *RL* network that has a time constant of $75 \mu\text{sec}$.

Multiplex Transmission. Two or more programs can be transmitted simultaneously on f-m broadcast transmitters. In this case, only the frequency-division method is practical. The additional channel may be used for facsimile, stereophonic sound, or special advertising to directed locations.

When facsimile is transmitted, the subcarrier may be either f-m or a-m. If amplitude modulation is applied to the subcarrier, the frequency must be 25 kc, and the sidebands must not extend more than 3 kc on either side of the subcarrier. Only 5 per cent modulation of the main carrier by the subcarrier is allowed for facsimile. Cross modulation of the subcarrier into the audio band must not exceed a level which is 60 db below 100 per cent modulation.

For transmission of anything besides facsimile, only frequency modulation of the subcarrier is allowed. The subcarriers are not specified, but the highest frequency transmitted must not exceed 75 kc, and the total modulation due to the subcarriers must not exceed 30 per cent.

Addition of the subcarrier signal to the modulated carrier could be accomplished at the modulator input terminals provided that the modulator meets the frequency response and intermodulation requirements. Few modulators can do this, however, and it is common practice to multiply the carrier that is modulated by the 50- to 15,000-cps signal, up to 8 to 10 Mc, and then to insert the subcarrier through a balanced phase modulator (60). A block diagram of such an exciter is shown in Fig. 12-4.

12-9. Television Transmitters. Visual transmitters are characterized by their broadband amplifiers with a modulation capability up to 4.5 Mc. Amplitude modulation is used because this type of modulation permits the use of the *vestigial-sideband* method of transmission. It is possible

to transmit a video signal having a bandwidth of 4.2 Mc and an aural signal having a bandwidth of 200 kc within a total bandwidth of 6 Mc. If frequency modulation were used in the visual transmitter, a total of at least 9 Mc would be required.

Figure 12-5 shows the idealized picture-transmission amplitude characteristic. The rapid drop below 0.5 Mc is accomplished by r-f filtering

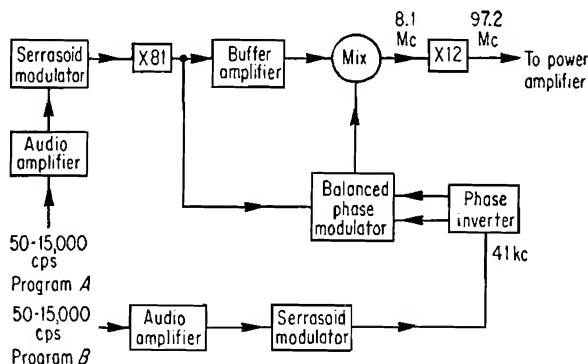


FIG. 12-4. Multiplex f-m exciter. (Courtesy of Electronics.)

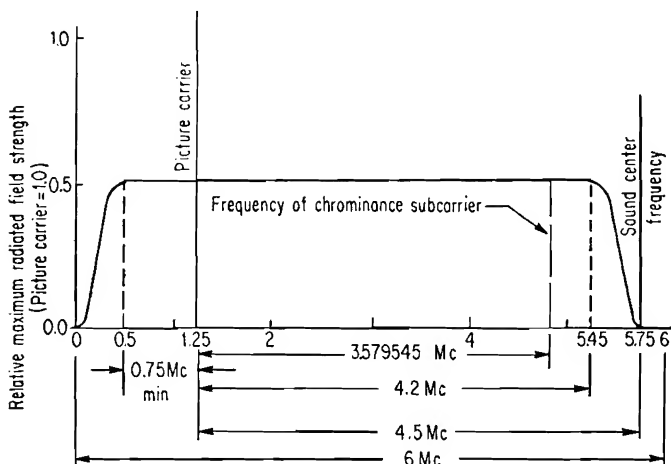


FIG. 12-5. Idealized picture transmission amplitude characteristic.

following the amplifier circuits or by the amplifier circuit characteristics. The rapid drop above 5.45 Mc is achieved by a low-pass filter before the video modulator.

The detailed monochrome and color waveforms will not be discussed as these are not generated in the transmitter (61).

D-C Transmission. One of the principal differences between an audio and a video modulator is that a video modulator can effectively transmit

a d-c component. Figure 12-6 shows the typical situation in an a-m aural transmitter. The area of the positive portion *A* will always equal

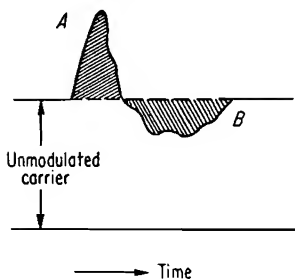


FIG. 12-6. Typical aural modulated wave (a-m).

the area of the negative portion *B*, although the amplitudes of the positive and negative peaks may be different. The average value of the transmitter output voltage remains constant over a period of time. In an a-m transmitter the average plate current to the modulated amplifier remains constant, and in an f-m transmitter the average frequency remains constant.

In the case of picture transmission, the average brightness of successive scenes is far from constant, and the picture may, in fact, be nearly all white or black for fairly long periods. There are other important reasons for d-c transmission:

1. A given transmitter can radiate about three times as much total power over a given period of time without increasing the interference range.
2. Constant sync peak level permits simple automatic gain control (agc) circuits to be used in the receiver.
3. Approximately 50 per cent more peak-to-peak swing from sync peak to white level is possible in the transmitter.
4. The modulator and modulated amplifier can be operated into the nonlinear region of the amplifier characteristics because the sync pulse can be easily prelengthened to take care of any compression.

The advantages of picture transmission with and without clamping are illustrated in Fig. 12-7 (62). The waveforms corresponding to a vertical white line on a black background and a vertical black line on a white background are shown for the condition without d-c transmission (without clamping) and with d-c transmission. When the areas above and below the unmodulated carrier condition are allowed to remain the same, it is seen that the transmitter output at the sync peak is less for the black background than for the white background. When the "back porch" of the sync pulse is held or *clamped* to a certain transmitter power output, the peak of the sync pulses will correspond to maximum power output regardless of the picture content. The average plate current to the modulated stages in a visual transmitter will increase as the black area of the picture is increased.

Clamping Circuits. The sync pulse may be clamped at the back porch or at the peak of the sync. Figure 12-8 shows a typical clamping circuit. Suppose that it is desired to grid-modulate an r-f amplifier which delivers the proper amount of black level power when the grid bias is -170 volts.

This bias voltage is also the plate voltage of the final modulator stage and, when switch S is closed, the bias on the final modulator stage is adjusted so that -170 volts is obtained at the plate. It is now necessary to arrange that switch S be closed *only* during the period when the video waveform is at the clamping level. The time constant of R and C must

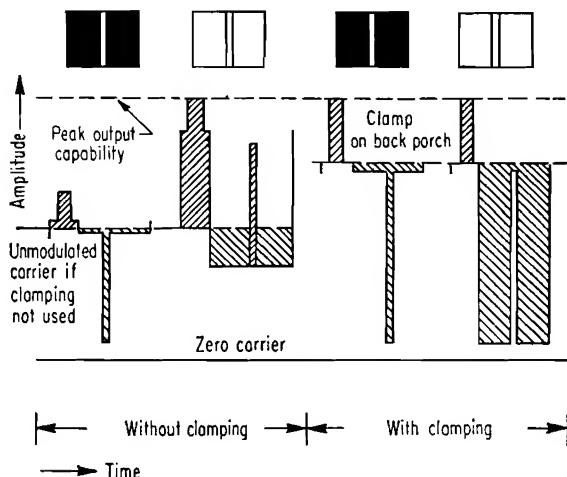


FIG. 12-7. Picture transmission with and without clamping. (Courtesy of RCA.)

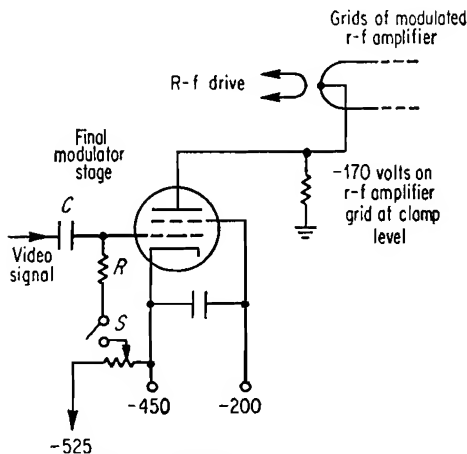


FIG. 12-8. Video modulator.

be small enough so that the charge on the capacitor and consequently the grid voltage reaches the correct value before switch S opens again. During the period that S is open, the leakage across it must be small enough so that C maintains its charge until the next time S closes, which will be at the end of the next horizontal line.

Figure 12-9 shows a switching circuit which is called a *keyed clamp*. Keying pulses of opposite polarity with respect to ground are applied to two diodes which are also connected with opposite polarities. The keying pulses are shorter than the length of the back porch and are generated with a blocking oscillator which has been delayed so that the pulses will be approximately centered on the back porch. Both diodes will conduct simultaneously and, if the circuit is balanced, the keying pulses will cancel out and will not be superimposed on the back porch of the video waveform.

Video Modulators. Video modulators must pass up to 3.5 Mc for monochrome service and 4.3 Mc for color service. In general, the high-frequency response is improved by reducing shunting capacities, reducing

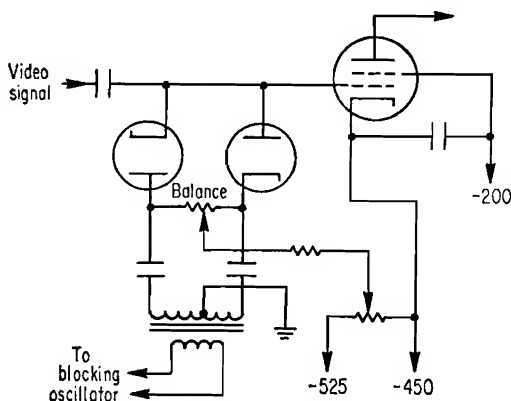


FIG. 12-9. Clamping circuit.

plate-load resistors and using series or shunt peaking circuits. It is also important to have good *transient* response in the modulator. The transient response of an amplifier is determined not only by the frequency response but also by the *phase* response (see Sec. 6-7).

The transient response of a transmitter is not generally specified because of the difficulty of performing tests and interpreting the results. The transient response of an amplifier can be examined by observing the output with a wideband oscilloscope when a square wave is applied to the input. Various definitions which are applicable to the transient response are illustrated in Fig. 12-10. This is the type of response obtained when the square-wave frequency is 100 kc or more. The over-all transient response of a transmitter is difficult to measure because it is not easy to obtain good transient response in the demodulation equipment. The following figures are representative of good transient response in a television system:

Rise time (10 to 90 per cent): 0.13 μ sec
 Anticipatory undershoot: 10 per cent
 First positive overshoot: 10 per cent
 First negative overshoot: 6 per cent
 Fourth overshoot and above: less than 2 per cent

The number of overshoot oscillations per second is called the *ringing frequency*.

When a low-frequency square wave is applied (60 cps or less) a *sag* or *tilt* may be observed in the output, and this should be less than 2 per cent for a 60-cps signal in a video amplifier.

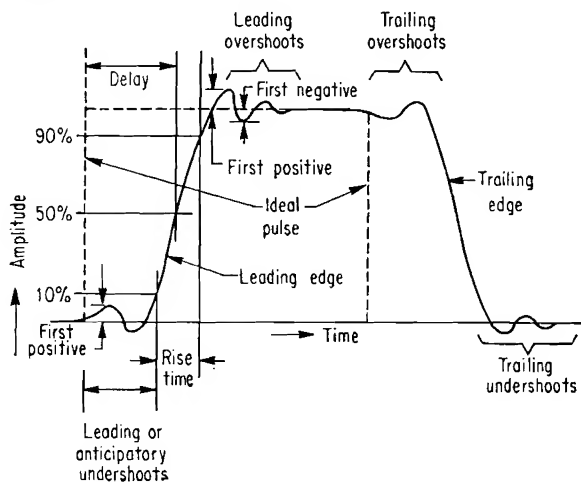


FIG. 12-10. Pulse characteristics.

It is found in practice that the best television pictures are produced by amplifiers that have a transient response which is symmetrical about the 50 per cent amplitude point (63). This can be accomplished by the use of special circuits known as *all-pass* phase-correcting filters which will affect the phase and transient response without changing the frequency response (64,65).

Color Transmission. A television transmitter must pass the color signal without distortion, and the possible effects of the transmitter characteristics on the signal will be discussed briefly. The color information is sent on a subcarrier whose frequency is 3.579545 Mc. If a simple band-elimination filter tuned to this frequency is inserted in the incoming video line to a transmitter, the color signal will be eliminated without disturbing the black-and-white (monochrome) picture.

The method of generating the color signal will not be discussed, but the resulting correct amplitudes and phase angles are shown in Figs. 12-11 and 12-12.

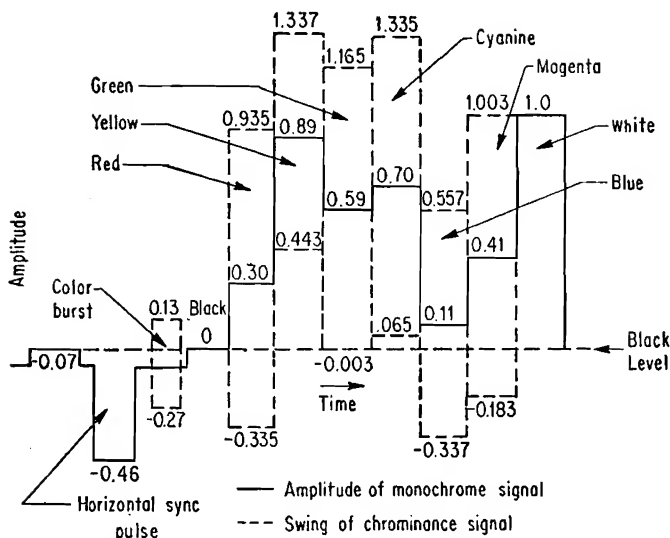


FIG. 12-11. Monochrome and chrominance waveform for saturated colors. (Courtesy of IRE.)

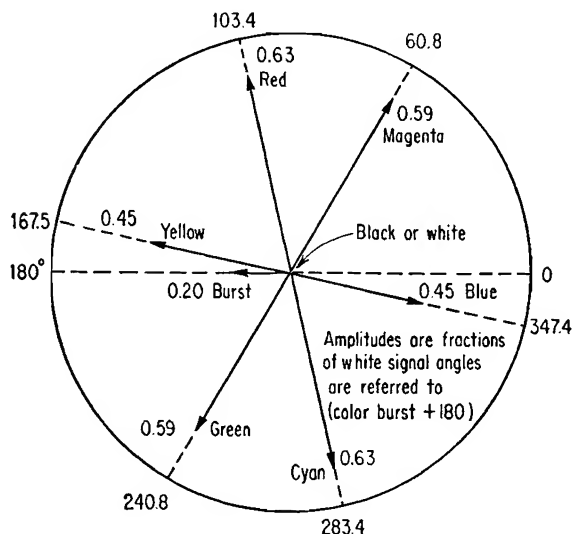


FIG. 12-12. Amplitudes and phases of color subcarrier for saturated colors. (Courtesy of IRE.)

Figure 12-11 shows the display that would be seen on a waveform monitor with saturated colors being fed in from a color bar generator. The solid line represents the monochrome amplitude corresponding to each color and the dotted lines indicate the swing of the subcarrier or chrominance signal. It will be noted that the monochrome amplitudes

corresponding to red, green, and blue will add up to 1.0 which corresponds to white.

Figure 12-12 shows the display that would be seen on a *vectorscope*, and the relative amplitudes and phases of the subcarrier can be seen. It will be noted that the horizontal and vertical projections of the red, green, and blue vectors add up to zero. White is actually obtained by simultaneously sending saturated red, green, and blue. The vector lengths shown in Fig. 12-12 are fractions of the white amplitude, and the angles shown are referred to the color burst plus 180° .

Effect of Transmitter Characteristics on Color. The following modulator and amplifier characteristics are of importance in color transmission (66):

1. *Amplitude vs. frequency response.* For monochrome transmission the FCC allows 12 db attenuation at 3.5 Mc; but for color the response must be not more than 2 db down at 4.18 Mc.

2. *Attenuation of color burst by clamps.* If the keyed clamp circuits operate during the "back-porch" interval, it is important that the clamp circuits do not attenuate the subcarrier. Insertion of a parallel resonant circuit tuned to 3.58 Mc between the clamping circuits and the grids being clamped will usually prevent this trouble.

3. *Differential linearity.* The output-vs.-input voltage characteristic or *linearity* of the modulator is important in monochrome transmission to obtain a satisfactory *gray scale*. As can be seen from Fig. 12-11, the transmission of the correct colors depends on the monochrome signal (solid line), and the linearity will have an effect on the colors. Differential linearity refers to a measurement of the subcarrier amplitude as the average value or monochrome level is changed (see Chap. 13 for methods of measurement).

4. Compression of the subcarrier amplitude will usually occur as the white region is approached, and the condition is corrected by means of *stretch* circuits. One method is to shunt the video signal with diodes which are biased to conduct in the midamplitude range and to cut off in the regions where compression is occurring.

5. *Differential phase.* The phase of the subcarrier relative to the color burst must not change as the monochrome signal amplitude changes (67). Poor differential phase characteristics are caused by diagonal clipping at the top of the subcarrier waveform, multiple transmission paths through r-f amplifiers, and variable impedance elements.

6. *Envelope delay.* If the chrominance information reaches the receiver picture tube at a different time than the monochrome information, the colors may be displaced horizontally and may not exactly overlay the black and white outlines. Typical receivers have a delay variation between the low and high frequencies, and the FCC has specified a

definite compensating delay in the color transmitter. This requires a circuit in or preceding the video modulator which introduces a constant envelope delay between 0.1 Mc and 3.0 Mc and then linearly decreases the delay so that it is $0.17\ \mu\text{sec}$ less at 3.58 Mc than it is at frequencies below 3.0 Mc ($0.05\ \mu\text{sec}$ corresponds to about one half of a picture element in the horizontal direction).

Figure 12-13 illustrates the meaning of envelope delay (68). The phase angle between the video input to a transmitter and the demodulated video from the output is shown plotted against the video frequency. If the line is straight, the equipment is said to have a *linear phase* characteristic, and the *time* delay will be consequently constant for all frequencies with no effect on the picture. If the time delay is greater, the

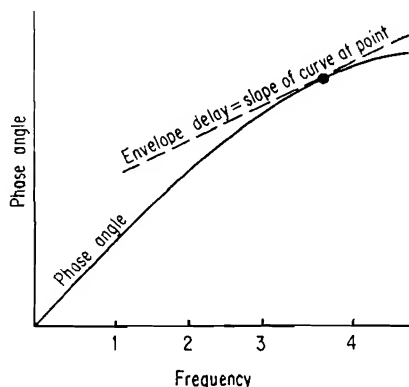


FIG. 12-13. Envelope delay.

line will have a steeper slope. If, however, the line starts to curve at the higher frequency end, it means that the delay is changing at the high end.

It can be seen from Fig. 12-13 that the difference in the phase angle between 3.0 and 3.5 Mc may be small, and it is always difficult to measure small differences between large quantities. A measurement of the *slope* of the phase-frequency characteristic at any frequency of interest may be easier, and this slope is called the *envelope delay*. (Measurement methods are discussed in Chap. 13.) Correction is achieved with delay equalizers which change the phase characteristic without affecting the frequency response. (See Refs. 64 and 65 for discussions of these circuits.)

12-10. Broadcast Repeaters and Translators. The service area of a broadcast station may be increased by the use of additional transmitters which are located some distance from the main transmitter and which rebroadcast the signal. The additional transmitters may be on the same frequency but it is more common to *translate* the signal to a different frequency.

Typical television translators in the United States operate on about 1-mv input on any channel between 2 and 13 and retransmit the signal at a power level of about 10 watts on a uhf channel between 70 and 83. The signal is not demodulated but merely converted to a new frequency by mixing with a crystal-controlled local oscillator. The translator is identified by periodically keying a Morse code signal.

12-11. R-F Generation for Noncommunication Purposes. There are a number of applications where relatively high levels of r-f energy are required, and these equipments closely resemble fixed station transmitters.

Induction or *dielectric* heating is used in manufacturing processes where the temperature must be closely controlled or where the heating must be localized to a small area of the part (69,70). Radio-frequency energy is also used for therapeutic purposes in medicine (71). The equipments used are called *diathermy* machines.

Another important use of high r-f powers is to drive nuclear research instruments. The *linear accelerator* and *synchrocyclotron* are examples (72-75).

To avoid interference with communications, these equipments must operate on industrial, scientific, and medical (ISM) frequencies.¹

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¹ 13.56, 27.12, 40.68, 915, 2,450, 5,850 and 18,000 Mc.

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TRANSMITTER MEASUREMENT TECHNIQUES

13-1. Transmitter Measurements. Performance measurements on a transmitter are important to ensure compliance with government regulations and are required for submission with license applications for certain classes of stations. Periodic checks will often reveal component deterioration which will allow replacement to be made before actual failure occurs. When a transmitter failure does occur, tests must be rapidly made to determine the cause of the trouble.

There is a considerable body of literature on the subject of electrical measurements. A number of books deal with basic measurement principles (1-9) while others discuss actual instruments (10-14). There are manuals which were written as guides for laboratory courses (15,16). A number of works are devoted to r-f and microwave measurements (17-23). Systems measurements are covered in sections of handbooks and general treatises (24-30) and also in a number of government (31-32) and industrial publications (33-36).

In this chapter typical tests will be outlined which may be used for determining component or equipment performance at the operating site. It is usually only necessary to determine whether or not a specified performance is being attained; a high degree of accuracy is not needed. The types of instruments that may be used to perform a test will be given, but details regarding the internal instrument circuitry will generally not be given. A listing of particular instruments that could perform each test would be impractical. A catalogue is available which shows the characteristics of most of the electronic instruments manufactured in the United States (37).

13-2. Power Measurements. The r-f output power from a transmitter is one of the principal indicators of its performance. The over-all efficiency from the primary power source to the transmitter output is also important where electric power costs are an appreciable percentage of the operating costs, and it may be necessary to know the total input power.

Primary Power Check. The input power may be most easily obtained by measuring the volt-amperes and multiplying by the *power factor*.

The power factor depends on such things as reactive current taken by transformers and motors and is unlikely to change unless the circuit is changed. If it is desired to measure the power factor, the power must be measured with a wattmeter. Then the power factor is given by

$$PF = \frac{\text{watts}}{\text{volts} \times \text{amperes}} \quad (13-1)$$

In a three-phase circuit the input power is given by

$$\text{Power} = 1.73 \times PF \times \text{line-to-line volts} \times \text{average line current} \quad (13-2)$$

If the equipment does not take any power from a line-to-neutral connection, the total power may be taken by the two-wattmeter method shown

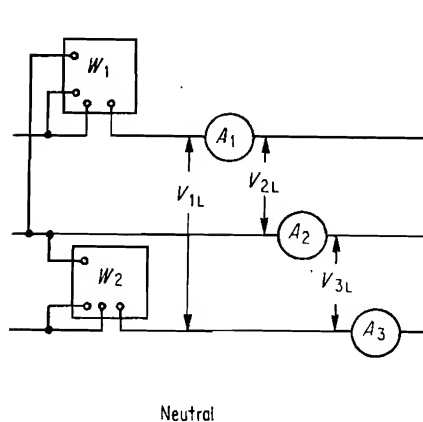


FIG. 13-1. Two-wattmeter method of measuring three-phase power.

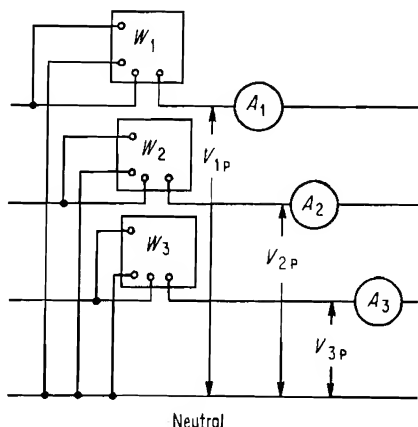


FIG. 13-2. Three-wattmeter method of measuring three-phase power.

in Fig. 13-1. The total volt-amperes will be given by

$$VA = \frac{V_1 A_1 + V_2 A_2 + V_3 A_3}{1.73} \quad (13-3)$$

The power factor will then be given by

$$PF = \frac{W_1 + W_2}{VA} \quad (13-4)$$

If, however, the neutral is connected and carries current, the three-wattmeter method, as shown in Fig. 13-2, must be used. The total power will then be

$$\text{Total power} = W_1 + W_2 + W_3 \quad (13-5)$$

and the total volt-amperes will be

$$VA = V_{1P}A_1 + V_{2P}A_2 + V_{3P}A_3 \quad (13-6)$$

and the power factor will be

$$PF = \frac{\text{total watts}}{\text{total volt-amperes}} \quad (13-7)$$

Note that in the two-wattmeter method the voltage coils of the wattmeter and the voltmeters are connected from line to line, while in the three-wattmeter method they are connected from line to neutral.

It is important to note that the method shown in Fig. 13-2 is only valid if the power company has connected the distribution transformers in a wye connection as shown in Fig. 13-3a. The power company will frequently choose the delta connection shown in Fig. 13-3b because the two untapped transformers can be cheaper units. In this case the three-phase and single-phase powers and power factors should be measured separately.

Audio-frequency Power. Measurements in the 20- to 15,000-cps range are made by measuring the a-c voltage across a resistor which simulates the load. At program levels a vu (volume units) meter is used. Zero vu corresponds to 1 mw of test tone in the resistor which terminates the line the meter is bridging. Vu also implies, however, that a particular meter movement is used.

Audio-frequency powers up to 100 watts can be read directly in dbm or watts on an output power meter. A typical instrument reads from 0.2 mw to 100 watts, and the terminating impedance can be adjusted from 2.5 to 20,000 ohms. The indicating meter is calibrated in rms values for a sinusoidal applied voltage, but is not a true rms indicating device, and there will be some error with nonsinusoidal voltages. An instrument which uses a thermocouple as the sensitive element indicates

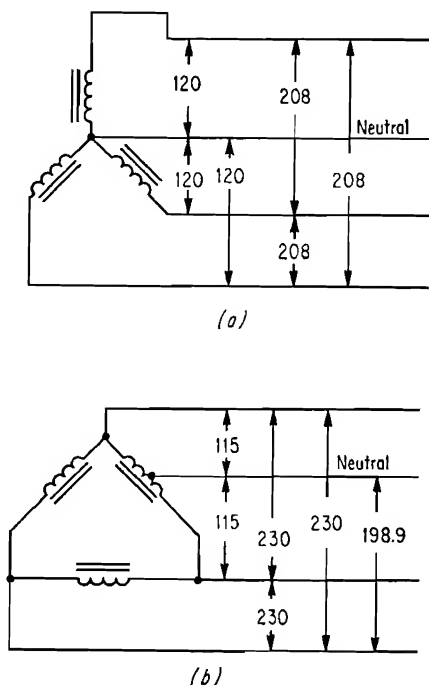


FIG. 13-3. Common methods of connecting transformer secondaries in power distribution circuits. (a) Wye connection; (b) delta connection.

true rms values. For a complete discussion of rms, peak-reading, and average-reading meters see Partridge (12). For higher a-f powers the a-f amplifier should be terminated in its correct load impedance, and either the current through the resistor or the voltage across it can be used to determine the power. The power is given by

$$P_{\text{a}} = I^2 R_L = \frac{E^2}{R_L} \quad \text{watts} \quad (13-8)$$

where I is in amperes and E is in volts. If a modulated r-f amplifier is to be simulated by a resistor, the value is given by

$$R_L = \frac{E_b}{I_b} \quad \text{ohms} \quad (13-9)$$

where E_b = plate voltage of modulated amplifier, volts

I_b = plate current of modulated amplifier, amp

The output voltage from a modulator may exceed the rating of available meters, and in this case the load resistance should be divided into two parts as shown in Fig. 13-4. The output power is then given by

$$\text{Output power} = \frac{E^2(R_1 + R_2)}{R_2^2} \quad \text{watts} \quad (13-10)$$

where E is in volts and R_1 and R_2 in ohms.

R-F Power up to 30 Mc. Radio-frequency ammeters of the thermocouple type are accurate within 2 per cent up to frequencies well above 30 Mc. Transmitter output powers are determined by taking the square

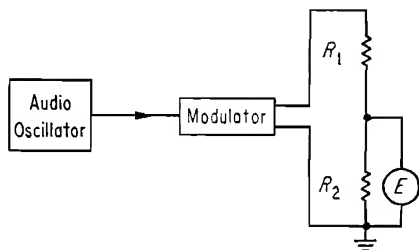


FIG. 13-4. Measurement of modulator output power.

of the antenna current and multiplying by the value of the antenna resistance at the point where the current is measured. Methods of determining the antenna resistance will be discussed later.

In case the operating frequency is changed or the antenna is changed, the FCC permits the following *indirect* method to be used for estimating transmitter

output power. This applies to broadcasting stations in the 0.54- to 1.6-Mc band.

$$\text{Operating power} = E_b I_b F \quad (13-11)$$

where E_b = plate voltage applied to final amplifier, volts

I_b = plate current taken by final amplifier, amp

F = factor to be obtained from Table 13-1

TABLE 13-1. INDIRECT METHOD OF DETERMINING TRANSMITTER OUTPUT POWER

Factor L	Method of modulation	Maximum rated carrier power, kw	Class of amplifier
0.70	Plate	0.1-1.0 kw	B BC
0.80	Plate	5 and over	
0.35	Low level	0.1 and over	
0.65	Low level	0.1 and over	
0.35	Grid	0.1 and over	

The 0.65 factor in Table 13-1 applies to all linear amplifier operation where efficiency approaches that of class C operation.

It should be pointed out that the power output of a c-w or a dsb a-m transmitter is defined as the *unmodulated* carrier power. If an rms (thermocouple) ammeter is used, the current value will increase by 22.5 per cent at 100 per cent modulation, but the d-c plate current to the final amplifier will not change.

SSB Power. In the case of single sideband there is usually no unmodulated power at all; in this case transmitters are generally rated in terms of *peak envelope power* (PEP). This is the r-f power at the peak of the audio cycle. It is equivalent to the average power in a c-w signal whose amplitude is equal to the peak of the modulated signal. If an ssb transmitter is modulated with a single tone, the PEP is equal to the average power; but this is not a recommended method of measuring the power output. The

reason is that many ssb transmitters will be overloaded with a sustained single tone. Another reason is that there is no clear indication of overload as there is when single-tone dsb modulation is checked. This will be discussed in more detail under modulation measurements. The recommended method of checking an ssb transmitter is to use two equal modulating tones applied at the same time. Measurements must be taken at a modulation level which is below that at which distortion occurs. Distortion is first indicated by a flattening of the audio peaks. The power can be measured by observing the output signal on an oscilloscope. This is shown in Fig. 13-5. If the scope is calibrated directly in volts then the power is given by

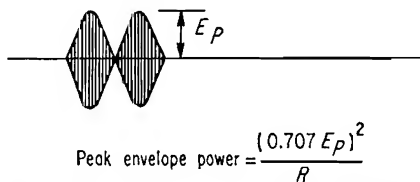


FIG. 13-5. Peak-envelope-power measurement with two-tone test.

at which distortion occurs. Distortion is first indicated by a flattening of the audio peaks. The power can be measured by observing the output signal on an oscilloscope. This is shown in Fig. 13-5. If the scope is calibrated directly in volts then the power is given by

$$\text{PEP}_0 = \frac{(0.707 E_p)^2}{R} \quad (13-12)$$

where E_p = peak voltage (half of peak-to-peak voltage)

R = load resistance, ohms

PEP_o = output peak envelope power, watts

It may not be convenient to calibrate the scope directly in r-f output volts; an alternative method is to obtain a calibration on the scope by operating the transmitter on continuous wave. This level will usually be somewhat lower than the PEP capability of the transmitter. The c-w power can then be measured by any of the methods described in this section. Then, with two-tone modulation the peak amplitude of the signal can be observed on the scope. The output power is then given by

$$\text{PEP}_o = \frac{E_p^2}{E_{\text{cal}}^2} \times P_{\text{cal}} \quad (13-13)$$

where E_p = peak amplitude on scope, in., boxes, or any arbitrary units

E_{cal} = observed amplitude on scope for calibrated power

P_{cal} = power at calibration, watts

PEP_o = output peak envelope power, watts

Another method is to observe the average power in watts during the two-tone test (38). If the two tones are pure sine waves and have equal amplitude, then the output power is

$$\text{PEP}_o = 2P_{\text{avg}} \quad (13-14)$$

This method must be used with some caution, however. A thermocouple-type ammeter and a bolometer bridge will read true power regardless of the applied waveform, but many other types of instrument will not.

Many transmitters are rated in terms of power input rather than output, and the relation between PEP input and d-c power input to a final amplifier is of interest. The PEP input power is now given by

$$\text{PEP}_{\text{in}} = 1.57 E_b I_b \quad (13-15)$$

where PEP_{in} = peak envelope power input, watts

E_b = plate voltage on final amplifier

I_b = plate current to final amplifier

The above relation is true under the following conditions:

1. Two tones are applied to modulator.
2. Each tone is sinusoidal.
3. Each tone has the same amplitude.
4. $I_b = 0$ when there is no modulation.
5. I_b is read on a meter that actually indicates average current under the two-tone condition.

As before, the method is dangerous because of uncertainty that condition 5 is being met. The current to the final stage can be measured with an

oscilloscope by measuring the voltage across a small series resistance. The input PEP is then simply given by

$$\text{PEP}_{\text{in}} = E_b I_p \quad (13-16)$$

where PEP_{in} = input peak envelope power, watts

E_b = plate voltage, volts

I_p = peak plate current, amp

The series resistance should be connected in series with the ground return to the power supply so that the scope will not be operated at an elevated potential. The scope can be calibrated by operating the transmitter on continuous wave and using the plate current meter as a reference.

Condition 4 above is not usually true because best linearity in a linear amplifier is usually obtained with some quiescent plate current. Then for the two-tone test with the other conditions still fulfilled, the input PEP is given by

$$\text{PEP}_{\text{in}} = 1.57 E_b I_b - 0.57 E_b I_q \quad (13-17)$$

where I_q = quiescent plate current, amp, and all other symbols are as before

The CCIR has recommended still another method of measuring ssb power. In this method the transmitter is modulated with two tones that produce sidebands of equal amplitude. The amplitude of these tones is adjusted until the intermodulation distortion just reaches -25 db. Then one of the tones is removed, and the output power is observed. The PEP will then be equal to four times the observed power with one tone.

The output *rating* of an ssb transmitter is dependent on what is specified as the allowable distortion, and it is important that the test method be always stated.

Dummy Antennas. In case the transmitter output must be checked without radiating a signal, a *dummy* or *artificial* antenna or load must be used. With standard broadcast transmitters, the dummy load impedance is usually equal to the transmission-line impedance before the antenna tuning unit. Such loads will have a resistance of 51 or 70 ohms and will consist of one or more noninductive resistors (39, 40). Loads for high-frequency broadcast transmitters will usually be balanced to ground and will be either 300 or 600 ohms. A delay-line type of load which is designed for use with rhombic antennas has a 40-kw capability (41). Air-cooled dummy antennas are available up to power levels of 5 kw, but at higher power levels, water cooling is more often used.

If the antenna tuning unit is to be fully checked the actual antenna impedance must be simulated. This means that a capacitor or inductor

must be connected in series with the antenna resistance which may go as low as 1 ohm. The series reactance must be capable of withstanding the high voltages that will be developed. For example, if 100 watts is to be developed in a 1-ohm antenna, 10 amp will be required; if the antenna reactance is 1,000 ohms, 10,000 volts will appear across the reactance. This is a typical problem with small-boat or other types of mobile antennas in the 2- to 3-Mc range. When a series reactance is

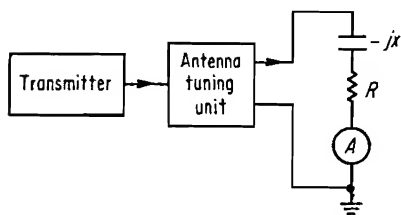


FIG. 13-6. Measurement of power in a low- or medium-frequency antenna.

used to simulate the antenna, the meter should be connected to ground as shown in Fig. 13-6.

At frequencies above 30 Mc the coaxial transmission line again becomes the common transmitter output connection. In addition, it is no longer possible to merely connect noninductive resistors in parallel to make a satisfactory load. Dummy antennas for this frequency range are made in coaxial form. A typical coaxial load resistor has an air-cooled rating of 1,500 watts and has a maximum variable standing-wave ratio of 1.1 up to 1,000 Mc when terminating a 50-ohm $3\frac{1}{8}$ -in. coaxial line.

Higher powers can be dissipated if water is circulated over the coaxial resistors. At frequencies above 1,000 Mc it becomes more satisfactory to dissipate r-f power directly in the water (42-44). This type of load is discussed in Chap. 11.

At frequencies above 30 Mc it is not practical to measure the current through the load resistor or the voltage across it. It becomes necessary to use directional couplers (45-49). These instruments are discussed in Chap. 11. Directional couplers directly calibrated in watts are available for frequency ranges of $2\frac{1}{2}$ to 1. "Termination" or "absorption" wattmeters are also available which consist of coaxial loads with built-in indicating meters.

Calorimetric Method. When a liquid-cooled dummy load is used, a basic measurement method may be applied. This depends on the law of conservation of energy and, providing there is no radiated heat loss, the power can be accurately determined from a measurement of liquid flow, input temperature, and output temperature. If the coolant is water (measured in gallons per minute), the power is given by

$$P_o = 0.264 \times \text{gpm} \times (T_{\text{out}} - T_{\text{in}}) \quad (13-18)$$

where P_o is in kilowatts and T in degrees centigrade, or

$$P_o = 0.147 \times \text{gpm} \times (T_{\text{out}} - T_{\text{in}}) \quad (13-19)$$

where P_o is in kilowatts and T in degrees Fahrenheit.

If the coolant is not water, the specific heat and also the specific gravity at the particular temperatures being used must be determined, and both these factors must be multiplied according to either of the above equations. If water is used and the flow is set to 3.79 gpm, the temperature difference in degrees centigrade will indicate the power in kilowatts directly. The specific heat and specific gravity for ethylene glycol at various temperatures is given in Figs. 13-7 and 13-8.

Bolometers. The bolometer is a small temperature-sensitive resistive element which is used to measure low power levels. The bolometer may be an instrument fuse, a *barretter*, or a *thermistor* and is mounted in a broadband impedance-matching mount. A *barretter* is a very fine wire, about 10^{-5} in. in diameter, usually platinum, and it has a positive temperature coefficient and a short time constant. The *thermistor* is a semiconductor usually composed of manganese and nickel oxides and has a negative temperature coefficient with a moderate time constant (50). Special instruments are available for measuring the change in bolometer resistance (51-53). Automatic self-balancing bridges give direct power

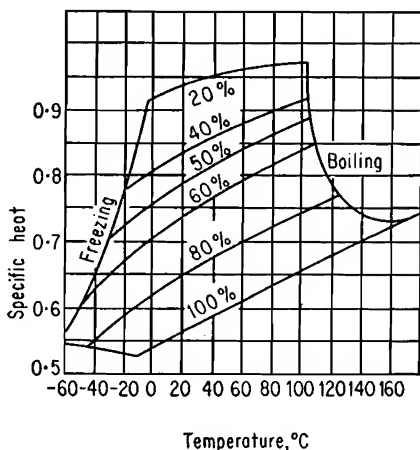


FIG. 13-7. Specific heat of ethylene glycol (antifreeze) and water mixtures. Percentages are by weight of glycol. (Copyright 1958 by Union Carbide Corp. Reprinted by permission of copyright owner.)

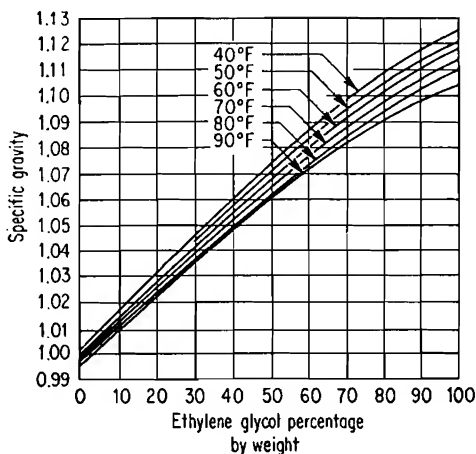


FIG. 13-8. Specific gravity of ethylene glycol. (Copyright 1958 by Union Carbide Corp. Reprinted by permission of copyright owner.)

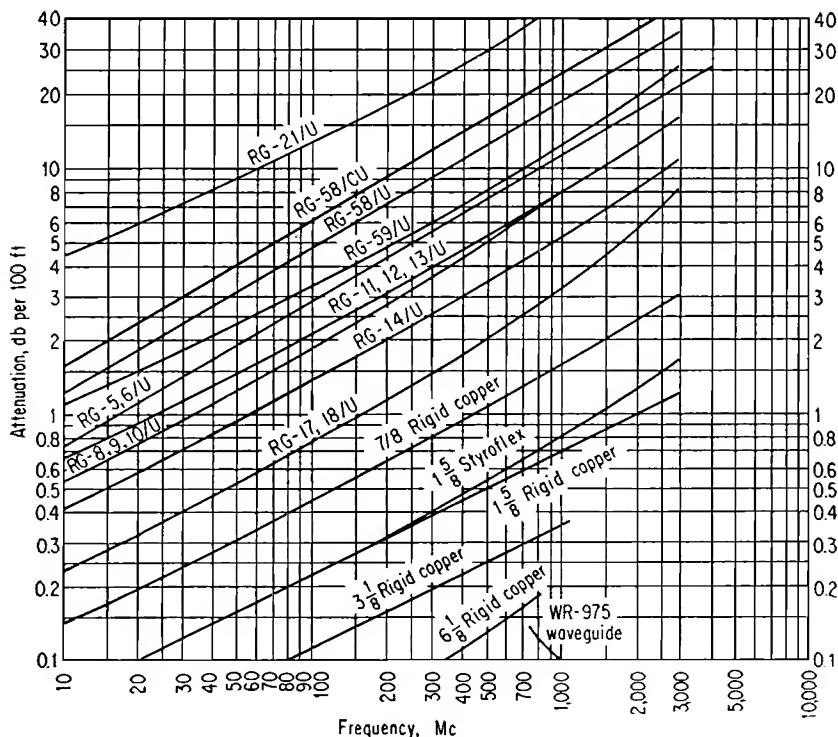


FIG. 13-9. Attenuation of coaxial cable. (Courtesy of Amphenol-Borg Electronics Corp. and Andrew Corp.)

readings from 0.02 to 10 mw after they have been balanced at zero r-f input power (54). Coaxial bolometer mounts are available for use between 10 and 10,000 Mc.

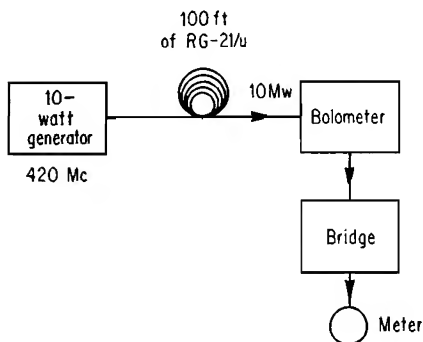


FIG. 13-10. Uhf power measurement using attenuating cable.

Waveguide bolometer mounts are available for use from 2.6 to 40 kMc. Power levels higher than 10 mw can be measured by feeding the power through attenuating cable of known attenuation and measuring the power at the end. Type RG-21/U is a typical attenuating cable. Attenuation-vs.-frequency graphs for a few types of coaxial cable are given in Fig. 13-9. The method is illustrated in Fig. 13-10. One hundred feet of RG-

21/U will have about 30 db of attenuation at 420 Mc, and the 10-watt level out of the generator will be reduced to 10 mw at the bolometer

input. It should be noted that a 1-db error in the estimation of cable attenuation will result in a 25 per cent error in the power measurement. The attenuation of the cable should therefore be measured, if possible, by checking it against some form of calibrated attenuator at the operating frequency.

Photometric Systems. A moderately accurate method of measuring power is illustrated in Fig. 13-11 and has been used as high as 420 Mc (55). The meter is calibrated by substituting a known source of direct current or 60-cycle power for the r-f power.

Self-balancing Calorimetric Meter. Direct-reading bolometer power meters read up to 10 mw while the direct-reading directional-coupler

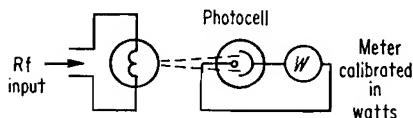


FIG. 13-11. Photometric method of measuring r-f power. (Courtesy of Electronics.)

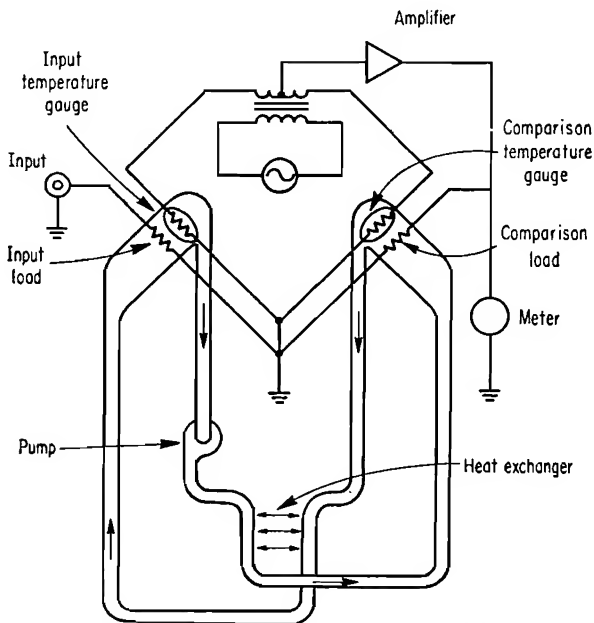


FIG. 13-12. Self-balancing calorimetric method of measuring r-f power. (Courtesy of Hewlett-Packard Co.)

meters cannot be read much below about 0.5 watt. A direct-reading calorimetric method is illustrated in Fig. 13-12. When power is fed into the input load, the instrument will automatically supply d-c power to a comparison load which will generate exactly the same amount of heat as that generated by the power fed into the unit (56). The amount

of power fed into the comparison load is then displayed on the meter. This method is used for direct reading of power in the range from 10 mw to 10 watts in one commercially available instrument.

Peak Power Measurements. Radar, TV, and other pulse transmitters are rated in terms of peak power. With a television transmitter the peak power is given by

$$\text{Peak power} = 1.68 \times \text{black-level power} \quad (13-20)$$

In other pulse transmitters the peak power can be calculated from a knowledge of the pulse shape and the duty cycle (57). Instruments are available which read peak power directly regardless of duty cycle and pulse shape for pulse widths between 0.25 and 10 μsec .

Miscellaneous Methods. There are a number of other power-measuring methods which will not be discussed in detail (58-62). Radio-frequency wattmeters have been built which sample current and voltage independently to provide a reading in watts (63). National Bureau of Standards Circular 536 is a comprehensive survey of r-f power measurements and gives over a hundred references including 20 American patents on the subject.

13-3. Frequency Measurements. It is important to know the output frequency of a radio transmitter. This is necessary to ensure compliance with government regulations. In the case of broadcasting stations, operation on frequency is important if generation of beat notes with other stations on the same frequency is to be avoided. In communication systems, it is important, of course, for the transmitted signal to lie within the pass band of the receiver. The efficiency of using the radio spectrum is very much dependent upon transmitter frequency tolerances.

Frequency and Time Standards. Frequency is the number of events per unit time. The unit of time in electronic work is usually the second. A *primary* standard of frequency is one whose frequency is determined directly in terms of time. Time is defined in terms of the earth's revolution around the sun. One second is the fraction $1/31,556,925.975$ of the tropical year 1900 (64). The *average* rotation of the earth relative to the sun is a mean solar day, and this is 86,400 sec. The *true* rotation of the earth is less, and this is called a *sidereal* day which is 86,164.091 sec. This is because the earth moves around the sun and will rotate once more than the number of solar days per year. A sidereal clock will agree with a conventional clock at about midnight September 21 of each year. A distant star will cross the meridian at the same time each day as indicated by a sidereal clock; this is one basis for checking a primary standard. Atomic standards which use the properties of the cesium atom or the ammonia molecule may also be used.

Primary frequency standards are maintained in various nations, and standardized radio frequencies are broadcast. By international agreement, standard frequencies are sent on the following channels: 2.5, 5, 10, 15, 20, and 25 Mc. MSF at Rugby, England, also sends a standard frequency on 60 kc. Time signals are sent by most of these stations (65). The Dominion Observatory in Ottawa, Canada, transmits voice-announced time signals each minute on 3,330, 7,335, and 14,760 kc.

WWV and WWVH. The U.S. National Bureau of Standards maintains standard-frequency stations at Beltsville, Maryland (WWV), and Hawaii (WWVH). The *frequency* transmitted by these stations is held as constant as possible with respect to the U.S. Frequency Standard which is based on atomic frequency standards at Boulder, Colorado. The *time* signals from WWV and WWVH are maintained in close agreement with a uniform time identified as UT-2 which is determined by the U.S. Naval Observatory. UT-2 is mean solar time corrected for polar variation and annual fluctuation in the rotation of the earth. The U.S. Frequency Standard was about 1.4 parts in 10^9 higher than a frequency based on the UT-2 second in 1957. On January 1, 1960, the U.S. Frequency Standard was changed to agree more closely with Ephemeris Time (ET) by establishing that the atomic oscillation of cesium is 9,192.631770 Mc and using this as a standard of frequency. The UT-2 second is not exactly constant, and every Wednesday at 1900 UT, step adjustments are made in the WWV time signals to make them agree with UT-2 time. Step adjustments are also made in the WWV frequency at this time if necessary. The changes do not exceed 1 part in 10^9 . WWVL at Sunset, Colorado, transmits c-w on 20 kc.

The WWV oscillator is stable to about 2 parts in 10^{10} per day. Random motion of the ionospheric layers prevents received radio signals from being observed with this degree of accuracy unless a long time average is taken. The National Bureau of Standards observes frequency data for 15 days preceding and 15 days following a given day and averages this to obtain a value for a given day. This is called the *30-day moving average* and is published monthly in the *Proceedings of the IRE*.

The time announcements from WWV and WWVH are given every 5 min in Morse code using the 0- to 24-hr system. The code time is Universal Time (same as Greenwich Mean Time or Greenwich Civil Time). At WWV only, a voice announcement of Eastern Standard Time is given before and after the code announcement.

Forecasts of radio propagation conditions are given twice an hour from both stations. A letter and a digit are used to classify conditions at the time of broadcast and for the next 6 hr. The digits from 1 to 9 indicate the forecast ranging from useless to excellent. The letters used are

N: Normal
U: Unsettled
W: Disturbed

Warnings of possible *geomagnetic* disturbances are also broadcast twice an hour when there is an unusually active solar region on the sun's disk. A succession of five letters in Morse code indicates the condition as follows:

AAAAA: State of alert

SSSSS: Special World Interval (SWI) will begin at 0001 UT the next day

TTTTT: SWI will finish at 2359 UT the same day

EEEE: State of no alert

Three extra long dashes instead of the five letters indicates that a Special World Interval is in progress. These signals are also sent by JJY in Tokyo and LOL in Buenos Aires.

WWV and WWVH transmit standard audio frequencies. One is 440 cps which is musical pitch for A above middle C. The other is 600 cps which can be conveniently divided down to either the 50- or 60-cps power-line frequencies. The second intervals are marked by pulses of 5 cycles of a 1,000-cps tone at WWV and 6 cycles of a 1,200-cps tone at WWVH. Minute intervals are marked by omission of the 59th pulse and by commencing the minute with two pulses 0.1 sec apart. A timing code is sent once per second for a 1-min duration 10 times per hour. The code is a 36-bit, 100-pps binary code; 9 binary groups per second include 2 groups for seconds, 2 groups for minutes, 2 groups for hours, and 3 groups for day of year. Blank spaces are available for additional information. Detailed characteristics of the two stations and the signals are given in Table 13-2. WWVL is keyed each 20 minutes.

It should be noted that, although the transmitted time intervals are accurate to within 1 part in $10^8 \pm 1 \mu\text{sec}$, random phase shifts or jitter due to atmospheric changes may cause errors up to 1,000 μsec . An interval of 24 hr should be used to compare a standard with WWV to within 1 part in 10^9 . Short-term errors of from -3 to $+3$ parts in 10^7 may be expected over a long-distance path owing to motion of the ionospheric layers. The Doppler effect will cause errors on a moving vehicle. This amounts to 1 part in 10^7 for a relative motion of 60 mph.

Except on 25 Mc, at WWV only, the tone signals are sent on upper-sideband full carrier. This is accomplished by using a separate transmitter which has an output power of about one-third that of the main transmitter. Voice announcements are made with conventional double sideband at WWV. The standard transmissions are changed from time to time, and details may be obtained from the National Bureau of Standards, Boulder Laboratories, Boulder, Colorado (66).

TABLE 13-2. CHARACTERISTICS AND SIGNALS OF STATIONS WWV AND WWVH*

Characteristic	WWV	WWVH
Radio frequencies, Mc, and powers, kw	2.5(1), 5(8), 10(9), 15(9), 20(1), 25(0.1)	5(2), 10(2), 15(2)
Frequency stability at transmitter	1 part in 10^9	5 parts in 10^9
Frequency deviation at transmitter	2 parts in 10^{10} per day	4 parts in 10^{10} per day
Seconds signal:		
Frequency.....	1,000 cps	1,200 cps
Duration.....	0.005 sec	0.005 sec
Accuracy.....	1 part in $10^8 \pm 1 \mu\text{sec}$	1 part in $10^8 \pm 1 \mu\text{sec}$
Audio tones:		
Frequency.....	440 and 600 cps	440 and 600 cps
Accuracy.....	1 part in 10^8	1 part in 10^8
Propagation forecast.....	19.5 and 49.5 min	9.4 and 39.4 min
Geomagnetic disturbance alert.....	4.3 and 34.3 min	14.4 and 44.4 min
Silent period.....	45-49 min	0-3, 15-18, 30-33, 45-48 min and 1900-1934 UT
440-cps tone.....	5-7, 15-17, 25-27, 35-37, 55-57 min	5-8, 25-28, 35-38, 55-58 min
600-cps tone.....	0-3, 10-12, 20-22, 30-32, 40-42, 50-52 min	10-13, 20-23, 40-43, 50-53 min
Timing code.....	7-8 min and every 5 min thereafter except 47-48	

* WWVL deviates less than 1 part in 10^{10} per day from a 20-kc signal which is -150 parts in 10^{10} below the U.S. Frequency Standard. Ultimate power is 300 kw.

TABLE 13-3. S-F BROADCASTS IN NATIONS OTHER THAN THE UNITED STATES

Call sign	Location	Carrier frequency, Mc	Modulation, cps	Power, watts
LOL	Buenos Aires, Argentina	2.5, 5, 10, 15, 20, 25	1, 440, 1,000	2,000
ZUO	Johannesburg, South Africa	5	1	100
ZLFS	Lower Hutt, New Zealand	2.5		30
	Moscow, USSR	10, 15	1	
MSF	Rugby, England	2.5, 5, 10, 60 kc	1, 1,000	500
JJY	Tokyo, Japan	2.5, 5, 10, 15	1, 1,000	1,000
IBF	Torino, Italy	5	1, 440, 1,000	300
	Uccle, Belgium	2.5		20

Other Standard-frequency Broadcasts. A partial list of s-f broadcasts in nations other than the United States is given in Table 13-3.

Locking an Oscillator to a Standard-frequency Station. If a strong signal is available from an s-f station, it is possible to lock an oscillator to the signal and to generate related frequencies (67). The block diagram of such a system is shown in Fig. 13-13. It is more usual to operate independent primary or secondary standards and compare them periodically with an s-f station. Crystal-controlled comparator receivers are available. The 20-kc WWVL signal is the best one to use.

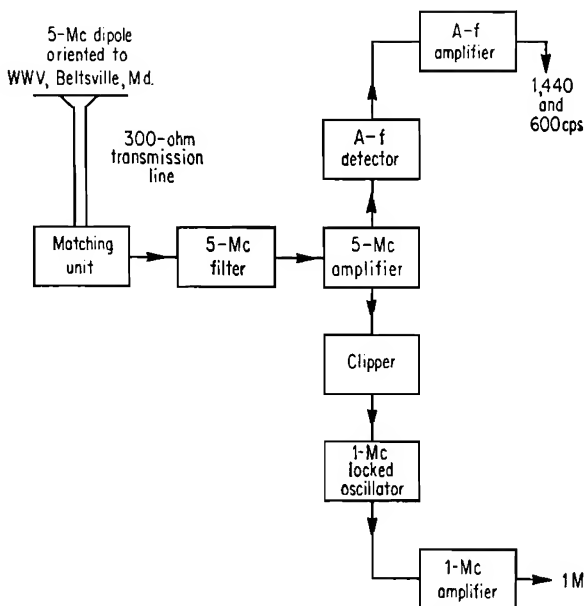


FIG. 13-13. Oscillator locked to standard-frequency station. (Courtesy of Electronics.)

Primary Standards. An s-f generator is defined as a *primary standard*, regardless of its stability, if it is used to control a clock. Quartz-crystal-controlled primary standards have been commercially available for many years. In a typical unit a 100-kc crystal is temperature-controlled to within 0.01°C. This frequency is divided successively to obtain 1 kc which is used to drive a 1-kc synchronous motor, which in turn, drives a clock.

There are also available primary frequency standards which are referred to the resonant frequency of cesium atoms, which is 9,192.631770 Mc (68). A typical unit generates output signals at 5 and 1 Mc and 100 kc.

Secondary Standards. A *secondary standard* is one whose frequency has been determined by comparison with another standard of frequency

regardless of the stability. In other words, if it has no clock, it is a secondary standard.

The crystal oscillators in typical secondary standards control frequency dividers which generate rectangular waves having a high harmonic content (69). It is these harmonics that are used for frequency measurements.

Much simpler units called *crystal calibrators* use one or two tubes or transistors and will generate harmonics of 50 or 100 kc up to about 30 Mc (70,71).

Frequency-measuring Stations. Broadcasting stations use frequency-deviation monitors and do not usually maintain secondary standards to

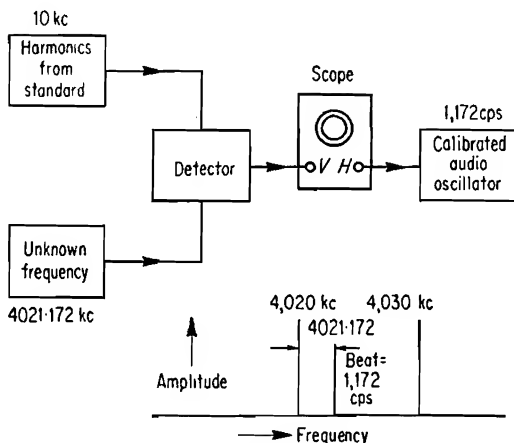


FIG. 13-14. Direct-beating method of frequency measurement.

directly check the monitor or transmitter. Frequency-measuring services are available to permit checks of the monitor accuracy to be made (72). The FCC and the Department of Transport in Canada also maintain frequency-monitoring stations (73).

Direct-beating Method. The direct-beating method is illustrated in Fig. 13-14. The unknown frequency together with harmonics from a primary or secondary standard are fed into a detector. A calibrated oscillator operating in the 0- to 5,000-cps range is connected to the horizontal amplifier, and the detector output is connected to the vertical amplifier of an oscilloscope. The oscillator is adjusted until a stationary circle appears on the oscilloscope. If the unknown frequency is 4021.172 kc, the audio oscillator will read 1,172 cps. Two important determinations must be made when using this method:

1. Which harmonic is being used?
2. Is the unknown above or below the standard?

Identification of the harmonic will be easy if an accurate calibration for

the detector is available. If not, the detector must be connected to the 100-kc signals or even to the 1-Mc signals to establish a reference on the detector dial, and then the harmonics may be counted until the unknown is reached.

If the unknown is not very close to the standard, the detector dial will reveal whether the unknown is above or below the standard. If the beat is only a few cycles, either the standard or the unknown must be moved a few cycles, and the direction and amount of shift should be known. In some secondary standards, a pushbutton is provided which lowers the frequency slightly. If the beat note goes down, the unknown is below the standard. If the beat note goes up, the unknown is above the standard, *provided* that the original beat note was more than half of the amount the standard was shifted. For example, suppose that the instrument is being used to check a frequency of 3,999.985 kc. The beat note will be 15 cps, and, when the test button is pushed, the beat note will go *up* to 25 cps. If the frequency was 4,000.015 kc, the beat note would again be 15 cps, and when the test button is pushed, it would go *up* again to 55 cps. Here a mere observation of the direction does not resolve the ambiguity. Table 13-4 shows how the unknown can be located.

TABLE 13-4. FINDING AN UNKNOWN FREQUENCY BY USING THE DIRECT-BEATING METHOD

If beat note moves . . .	Unknown Relative to Standard Is . . .	
	For standard moved down or unknown moved up by X cps	For standard moved up or unknown moved down by X cps
Up by X cps	High	Low
Down	Low	High
Up but was $X/2$ or less before moving	Low	High

There will also be an ambiguity when the beat note is close to 5,000 cps. If, for example, the beat is 4,990 cps and the detector is tuned to the unknown, there will also be a beat with the next harmonic of 5,010 cps. If the detector dial is adjusted so as to reduce the 5,010 without reducing the 4,990 beat, it should be possible to identify which harmonic is causing which beat. Another method is to observe the lower of the two beats and to shift the standard slightly. The position of the unknown can be found from Table 13-4.

When the frequency of a transmitter which is operating intermittently is to be measured, or when the signal is subject to fading, a *transfer oscillator*

may be used. This is adjusted to zero beat with the signal to be measured and then the transfer oscillator is measured. The transfer oscillators are also good harmonic generators and may be zero beat with signals up to 100 Mc. The fundamental frequency of the transfer oscillator is then accurately read.

Electronic Counters. Frequency may be read directly by means of an electronic counter which counts cycles during an accurately determined period of time (74-77). A typical counter has a range from 10 cps to 10 Mc and a selection of five periods from 0.001 to 10 sec. The periods are crystal-controlled, and the instrument has a short-time stability of 3 parts in 10^8 and a long term stability of 5 parts per 10^8 per week. Below 300 cps, more accurate readings can be obtained by using the unknown to operate the gate and using one of the internal standard frequencies to operate the counter. The unknown frequency will then be equal to the standard frequency used divided by what the counter reads. Frequency converters are available to extend the range of the instrument to 220 Mc. A typical counter is shown in Fig. 13-15.

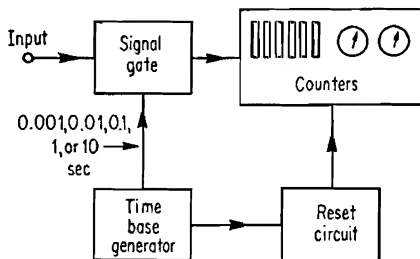


FIG. 13-15. Events-per-unit-time counter. (Courtesy of Hewlett-Packard Co.)

Frequency-deviation Monitors. For broadcast stations, the FCC requires that the frequency monitor be "continuously operated." The usual method of measurement is to set up a crystal oscillator which is operating at a frequency different from the assigned frequency. The beat frequency between the transmitter output and the standard crystal is then measured and displayed on a meter which is calibrated directly in cycles deviation from the assigned frequency. In monitors for the standard a-m broadcast band the difference between the standard crystal and the assigned frequency is exactly 1,000 cps when the standard crystal is exactly on frequency. The meter on this instrument reads ± 30 cps full-scale and is readable to within 1 cps. The FCC requires a range of at least ± 20 cps and readability to within 1 cps.

Monitors for the color-television subcarrier are similar to the standard a-m broadcast monitors. The FCC does not require continuous monitoring of this frequency but does require that it be held within ± 10.7 cps. The color subcarrier is 3.579545 Mc. A typical instrument does not operate on the television carrier but must be connected directly to the color subcarrier oscillator. Similar frequency-deviation monitors are available for the h-f range between 1.6 and 15 Mc.

Monitors for checking the deviation of the center frequency of an f-m

transmitter from the assigned carrier frequency must have the standard crystal offset by more than the peak swing of the modulated signal. The usual offset frequency is 150 kc.

There is some ambiguity concerning the term *frequency deviation* in an f-m transmitter. In the IRE Standards on Modulation Systems: Definitions of Terms, 1953, frequency deviation is defined as the peak difference between the instantaneous frequency of the modulated wave

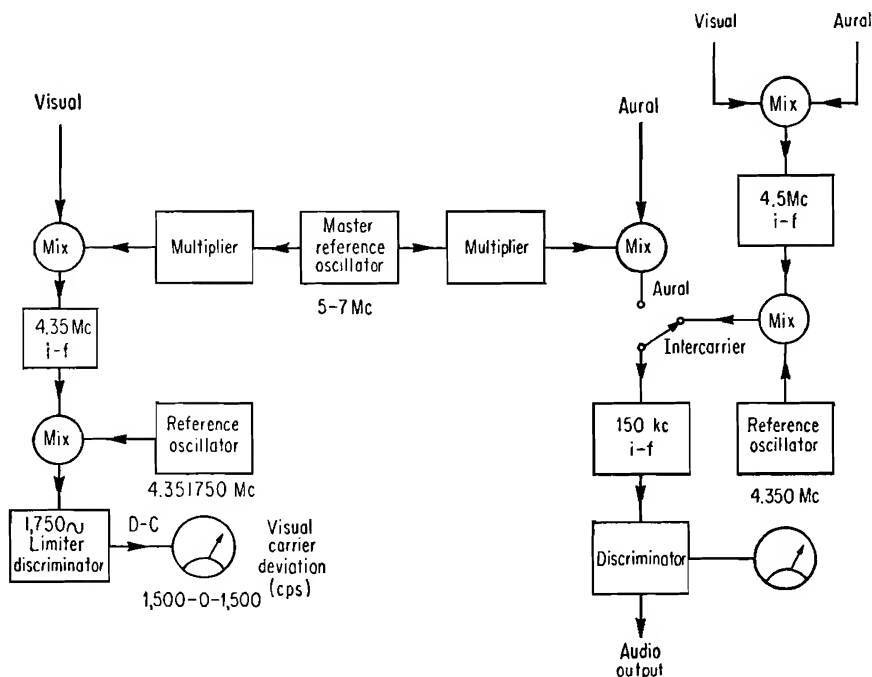


FIG. 13-16. Frequency-deviation monitor for TV transmitter. (Courtesy of General Radio Co.)

and the carrier frequency. The FCC, on the other hand, defines frequency *swing* as the instantaneous departure of the frequency of the emitted wave from the center frequency resulting from modulation. The FCC reserves the term *deviation* for the difference between the average frequency and the assigned frequency. The instrument manufacturers follow the FCC practice, and the meters marked with the word *deviation* show the frequency error.

In the case of a television transmitter, there are three frequencies of interest. These are the visual carrier, aural center frequency and the center frequency of the *intercarrier* or difference between the aural and visual carriers. In the case of a color-visual transmitter, a fourth frequency of interest is the color subcarrier.

Typical television monitors display the visual frequency at all times and either the aural or intercarrier center frequencies as selected by a switch. The elementary circuit of one type of monitor is given in Fig. 13-16. It is usual practice for the frequency monitor to include the aural-modulation monitor.

There is a possible ambiguity when measuring the visual frequency on this type of instrument. If the visual carrier is 3.5 kc high in the case of the unit shown in Fig. 13-16, the visual meter will show zero deviation. The intercarrier indicator would, however, indicate the error. The FCC requires another method of resolving this possible ambiguity. If the reference oscillator is moved down slightly with a test button, the visual deviation meter will also move down if the carrier is on the correct side of the reference oscillator but will move up if the carrier is on the wrong side.

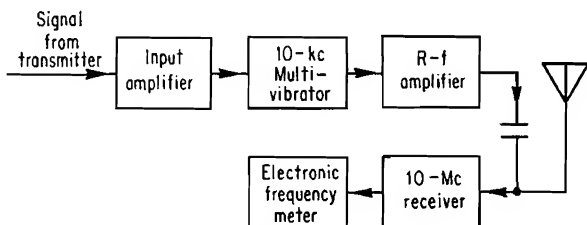


FIG. 13-17. Frequency measurement by beating harmonic of transmitter with standard station. (Courtesy of *Electronics*.)

Electronic Frequency Meters. The beat frequency between a standard oscillator and an unknown frequency or any unknown frequency in the range below 100 kc may be read on direct-reading frequency meters (78). An accurate measurement of a broadcast station frequency may be made by synchronizing a 10-kc multivibrator to the transmitter output and detecting the beat between a harmonic of the multivibrator and WWV on a communications receiver (79). The beat may be measured with a frequency meter. The high-low ambiguity must be resolved by shifting the transmitter slightly. The method is shown in Fig. 13-17. If, for example, the station frequency is 1,170 kc and the beat is measured against the 10-Mc WWV signal, the beat must be multiplied by 117/1,000 to obtain the transmitter deviation.

Transfer Oscillator Method. The method shown in Fig. 13-17 is only useful for the standard broadcast band. It turns out, however, that *all* broadcasting stations (except for international high frequency) are on even 10-kc frequencies. For example, the U.S. channel-10 visual frequency, when "offset" low, is 193.24 Mc. This can be accurately checked with the use of a crystal calibrator that generates 10-kc and 10-Mc harmonics and a transfer oscillator which operates in the 3- to

4-Mc range (80). The transfer oscillator need not be calibrated and only needs to be stable enough to stay within a few cycles for a few seconds. The equipment arrangement is shown in Fig. 13-18, and the principle is illustrated in Fig. 13-19. Crystal calibrators are suitable units and should be standardized against WWV before a measurement is made. The calibrator is first switched to the 10-Mc position, and the transfer oscillator is adjusted until a fairly small beat note is detected.

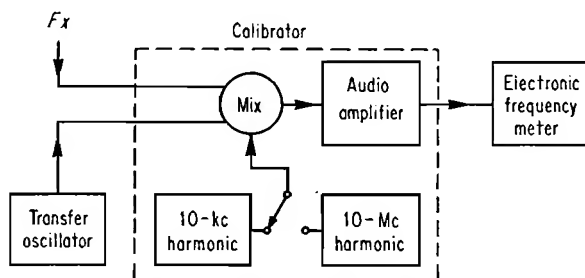


FIG. 13-18. Transfer oscillator method. (Courtesy of IRE.)

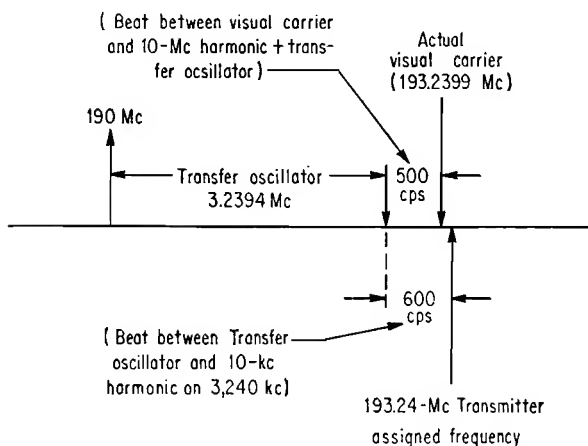


FIG. 13-19. Beats generated by transfer oscillator. (Courtesy of IRE.)

Then the calibrator is switched to 10 kc, and the beat note is measured. The *difference* between the two beat notes will be the difference between the unknown and an even 10-kc interval. It must be assumed that the TV or f-m station is at least within 10 kc of the proper frequency. As a final step, to resolve the high-low ambiguity, the transfer oscillator should be moved a small amount, and the check repeated. The difference should remain unchanged. If the difference does change, the transfer oscillator should be moved down through zero beat and up onto the other side of the unknown. The direction of the beat note change and

the direction the transfer oscillator is moved should then be observed. If moving the transfer oscillator *down* moves the beat note *up*, then the transfer frequency is *below* an even 10-kc harmonic. Now the position of the unknown relative to the even 10-kc can be visualized from Fig. 13-19. Remember that in the 10-Mc position the beat with the unknown is being measured, while in the 10-kc position the beat with the standard is being measured. The high-low ambiguity can be resolved from Table 13-5.

TABLE 13-5

If transfer oscillator moved . . .	and beat moves . . .	and 10-Mc relative to 10-kc beat is . . .	then unknown is . . .
Down	Up	Higher	High
Down	Down	Higher	Low

The procedure can be illustrated by the following example: A channel-10 offset low visual carrier is being measured, and the observations are shown in Table 13-6. The broadcaster is usually more interested

TABLE 13-6

If transfer oscillator moved . . .	and beat moves . . .	and 10-Mc beat is . . .	and 10 kc beat is . . .	differ- ence is . . .	and unknown is . . .
Down	Up	500	400	100	193.2401
Down	Up	500	600	100	193.2399
Down	Down	500	300	200	193.2398
Down	Down	500	700	200	193.2402

in getting the carrier exactly on frequency than in measuring the exact frequency. The procedure then is simply to adjust the transmitter until the beat note is the same with the calibrator on either the 10-Mc or 10-kc position. A check must be made, of course, to make sure that the beat note remains the same for the two positions with a slightly different transfer oscillator frequency. For this procedure an electronic frequency meter is not needed, and most observers can bring the two beat notes within a few cycles simply by ear with a headset.

Sliding Harmonic Method. Another method which is useful in the range from 100 to 3,000 Mc is to use an *interpolating* frequency standard (81). This is an accurately calibrated oscillator operating in the range of 1,000 to 1,010 kc, and it is connected to a 1-Mc multivibrator which generates harmonics up to 200 Mc. The harmonics and the unknown frequency are connected into a *heterodyne frequency meter* which operates

in the range of 100 to 200 mc. The procedure is to adjust the heterodyne frequency meter to zero beat with the unknown and then to adjust the interpolating standard to zero beat again. The calibration on the heterodyne frequency meter is accurate enough to identify which 1-Mc harmonic is used. The interpolating standard reading is then divided by

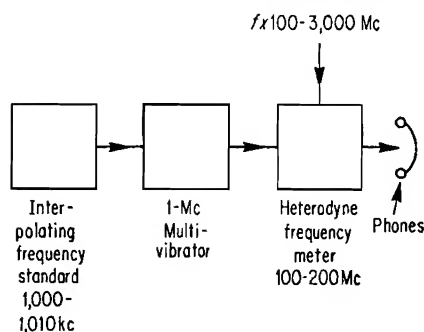


FIG. 13-20. Sliding harmonic method.
(Courtesy of IRE.)

the number of divisions per megacycle at the particular harmonic used, and this number is then added to the harmonic. The method is shown in Fig. 13-20. Harmonics of the heterodyne frequency meter are used for frequencies from 200 to 3,000 Mc. For example, suppose the unknown is 2,425 Mc. A beat will be heard on the heterodyne frequency meter at 121.25 Mc. The 121 can be read off the meter. If the interpolating oscillator shows 400 divisions per Mc for this harmonic, it will read 100 divisions when it is zero beat with the heterodyne frequency meter, whose frequency is then given as

$$121 + \frac{100}{400} = 121.25 \text{ Mc}$$

Frequency Measurement with Frequency Synthesizers. Frequency synthesizers (see Chap. 2) may be used for measurements by zero-beating the output or a harmonic with the unknown.

Direct-reading Frequency Meters. Up to this point the discussion has been about quite accurate measurements. There are many applications where such precision is not necessary, and the equipment may be much less expensive. Measurements within about 0.01 per cent can be obtained with direct-reading frequency meters such as the BC-221 and TS-175/U (82). These instruments may be used as signal generators to beat with the unknown in a receiver, or, if a strong signal is available, the internal detector in the instruments may be used.

These instruments must be used with caution when the frequency being measured is much higher than the fundamental oscillator frequency. Beats will be heard at more places across the dial as the unknown is increased. The question is: What multiplying factor should be used for any given reading? The unknown frequency can be determined by observing f_1 in terms of the oscillator *fundamental* and then observing the next higher beat f_2 (83). The unknown is then given by

$$f_x = \frac{f_1 f_2}{f_2 - f_1} \quad (13-21)$$

If the unknown has harmonic output, a weaker beat between f_1 and f_2 will be heard. The fundamental ranges of the instruments mentioned above are shown in Table 13-7. For example, if 75.5 Mc is being checked with the BC-221, beats will be heard at 3,973.684, 3,775, and 3,595.238 kc. The lowest frequency that can be used with this method is three times the bottom end of the fundamental range used.

TABLE 13-7. FREQUENCY RANGES OF TWO DIRECT-READING FREQUENCY METERS

Instrument	Ranges	
	Low	High
BC-221	125-250 kc	2-4 Mc
TS-175/U	80-200 Mc	

One basic principle to remember in using these meters is that it is always more accurate to heterodyne or *mix up* than to *multiply up* to the frequency being measured. Heterodyning is shown in Fig. 13-21. For example, if a frequency of 7,130 kc is to be measured the frequency meter will be set at 130 kc and will beat with the 70th harmonic of the 100-kc calibrator. Now, if the accuracy of the frequency meter is 1 part in 10,000 and the calibrator is set with WWV, the accuracy of the final reading will be 1 part in 700,000 or a 70-fold improvement in accuracy. This method also has ambiguities as beats will be heard in the receiver when the frequency meter is on 130, 170, and 230 kc. The rule is to ignore all beats between 225 and 250 and to also ignore all beats which move *down* on the receiver when the frequency meter is moved *up*.

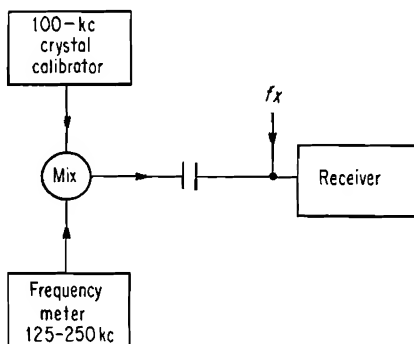


FIG. 13-21. Heterodyning direct-reading frequency meter with harmonic of crystal calibrator.

Wavemeters. The wavemeter is a rather crude frequency-measuring device which consists merely of a resonant circuit. Two general types are the *absorption* type, which absorbs some power from the circuit being measured, and the *reaction* type which will react in some way on the circuit being measured. The absorption types will be fitted with a detector and a meter to indicate resonance. The wavemeter has largely been replaced by the *grid-dip meter* on frequencies below 400 Mc. In the microwave region, cavities with very high Q 's can be made, and when

these are fitted with a micrometer adjustment, accuracies as good as 0.08 per cent can be achieved. In this range they are generally called *frequency meters*. Sections of waveguide fitted with reaction-type cavities to cover the range from 6,000 to 90,000 Mc are available. When the cavities are tuned to resonance, a dip will be observed in an output meter which terminates the waveguide.

Grid-dip Meters. The grid-dip meter is a very useful instrument which serves as either an absorption- or reaction-type wavemeter, but has much greater sensitivity than a simple wavemeter. A "cold" tuned circuit coupled to the grid-dip meter will give a "dip" in the grid current indicated as the meter is tuned through the resonant frequency of the coupled circuit. When the tuned circuit is "hot," the grid-dip meter is switched to the "diode" position, and the frequency being generated by the tuned circuit will be indicated by a maximum rather than a minimum reading on the grid-dip meter.

Typical instruments cover the range from 100 kc to 940 Mc. Portability of the instrument is improved by using transistors and battery

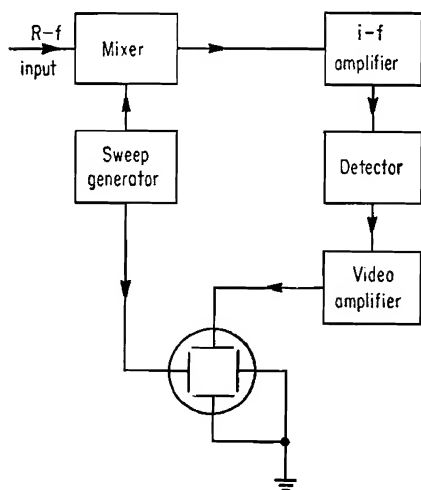


FIG. 13-22. Block diagram of spectrum analyzer.

operation, although in this case the use of the term "grid" becomes a misnomer (84). The instruments are fairly easy to build for operation at frequencies below 250 Mc. For higher frequencies special techniques are required (85).

The grid-dip meter is not an accurate frequency-measuring device but may be more useful for certain types of measurement than some of the methods mentioned earlier in this section. For example, the frequency of a parasitic oscillation in a transmitter may be easily identified with a grid-dip meter. The grid-dip meter is also very useful in setting up tuned

circuits in a transmitter before the application of power.

Spectrum Analyzers. A spectrum analyzer is essentially a narrow band superheterodyne receiver which is repeatedly swept in frequency over a selected portion of the r-f band. The output is displayed on an oscilloscope with the horizontal deflection synchronized with the frequency sweep. The vertical deflection indicates the amplitude of the signals (86-89). A basic block diagram is shown in Fig. 13-22. The i-f output of a communications receiver may be fed into an instrument

which is called a *panoramic adapter*. These give a similar display and are not difficult to construct (90). From 50 down to 0.5 cps it is more convenient to use an *oscillograph*, which is the same as an oscilloscope except that the "writing" is done with pen and ink on paper instead of a spot on a cathode-ray tube.

Spectrum analyzers are particularly useful for frequency-measuring purposes at super-high frequencies (91). The average commercial analyzer is calibrated within about 1 per cent accuracy. A more accurate reading can be obtained by feeding signals from an shf secondary standard (92,93) into the analyzer along with the unknown. The accuracy with which the unknown can be determined will depend on the horizontal linearity of the scope and also on the width of the i-f filters in the analyzer. A typical display of this kind is shown in Fig. 13-23. The spectrum analyzer provides a convenient method of calibrating cavity-type frequency meters as shown in Fig. 13-24. The pip on the scope will dip 10 to 20 per cent when the frequency meter is tuned to the frequency being observed.

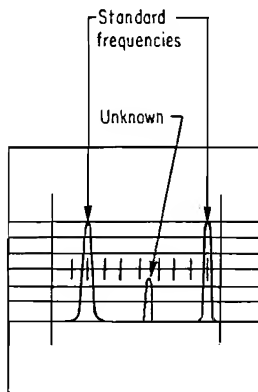


FIG. 13-23. Frequency measurement with spectrum analyzer. (Courtesy of Polarad Electronics Co.)

Miscellaneous Methods. There are a number of other methods of measuring frequency such as paramagnetic resonance absorption (94), interferometers (95,96), and others (97-100) which will not be discussed in detail.

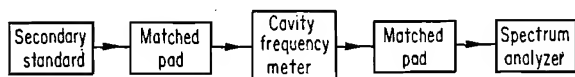


FIG. 13-24. Method of calibrating cavity-type frequency meter with spectrum analyzer. (Courtesy of Polarad Electronics Co.)

13-4. Time Measurements. The accurate measurement of short time intervals is important in testing pulse transmitters. The duration of the pulses as well as the interval between pulses must usually be measured. This is also a large subject, and it has been covered by at least one book (101).

One method of measuring short times is to calibrate the horizontal trace on an oscilloscope by means of signals from a secondary standard. For example, if a 1-Mc square wave is displayed on a scope, the leading edges will be 1 μ sec apart. Many oscilloscopes have an internal base calibrator. A typical unit has time markers spaced 0.05, 0.1, and 1.0 μ sec.

When accurate measurements must be made, pulses are obtained from an accurately calibrated external source such as a digital delay generator (102).

A direct numerical display without the use of an oscilloscope is also possible with an electronic counter. Instruments are available which will display intervals from $1\text{ }\mu\text{sec}$ up to 100 days.

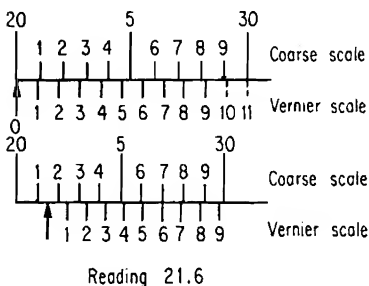


FIG. 13-25. Typical vernier scales.

Now the coarse scale could be represented by pips from a standard oscillator while the vernier scale could be pips from a slightly faster oscillator. If the timing pulses are less than 1 per cent of the spacing, better resolution can be obtained by making 100 vernier pulses equal to

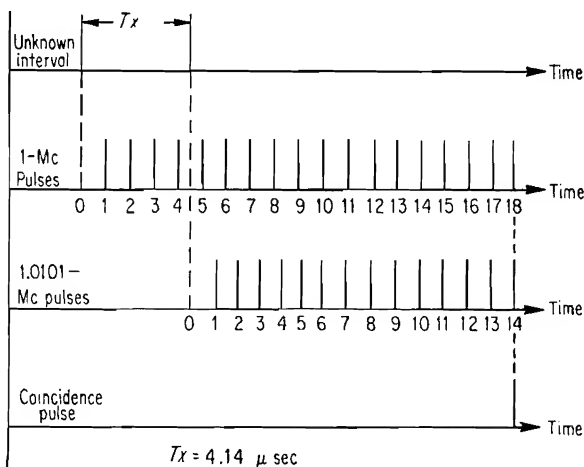


FIG. 13-26. Vernier principle for accurate time-interval measurements. (Courtesy of Electronic Inds.)

99 coarse pulses. The method is illustrated in Fig. 13-26. The first signal starts a counter which is counting 1-Mc pulses, and the second signal starts another counter which will count 1.0101-Mc pulses. The first time a coincidence occurs between the two sets of pulses, a coincidence circuit resets both counters. The time between the two pulses

is then given by

$$T_z = C - V + .01V \quad (13-22)$$

where T_z = time, μsec

C = reading of coarse counter

V = reading of vernier counter

13-5. A-M and F-M Measurements. When a transmitter is delivering the proper power output on the assigned frequency, it is safe to commence modulation. If the modulation is too low, the signal-to-noise ratio at the receiver may be degraded, while if overmodulation occurs, distortion will be introduced at the receiver. Excessive modulation may also cause interference to other transmitters. The FCC

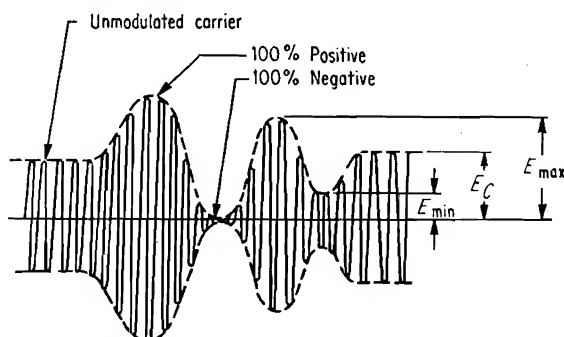


FIG. 13-27. Amplitude modulation percentage measurement.

requires that the percentage of modulation of broadcast transmitters be maintained at no less than 85 per cent and no more than 100 per cent on peaks.

Amplitude Modulation. Amplitude modulation is illustrated in Fig. 13-27. Percentage modulation is defined as follows:

$$\text{Positive \% modulation} = \frac{E_{\max} - E_c}{E_c} 100 \quad (13-23)$$

$$\text{Negative \% modulation} = \frac{E_c - E_{\min}}{E_c} 100 \quad (13-24)$$

$$\text{Average \% modulation} = \frac{E_{\max} - E_{\min}}{E_{\max} + E_{\min}} 100 \quad (13-25)$$

These values may be measured directly on an oscilloscope. Figure 13-28 gives a template that may be set on the scope face. The scope gain is adjusted so that the maximum peaks are at the upper and lower edges of the graph. The paper is then moved horizontally until the envelope minimum touches the curve, and the percentage modulation is read off the scale. An accurate reading at low modulation levels can be made by

making the modulation fill up the scope as shown in Fig. 13-29. A scope that deflects on direct current is preferred and the three points of interest can be accurately calibrated with an external d-c source.

The FCC requires the use of continuously operating monitors in broadcasting stations. These must be capable of being switched to read either positive or negative peaks. One commonly used instrument has

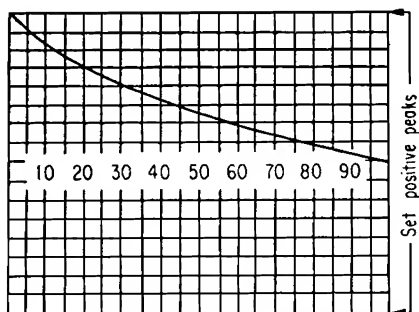


FIG. 13-28. Scale for a-m measurement.
(Courtesy of Electronic Inds.)

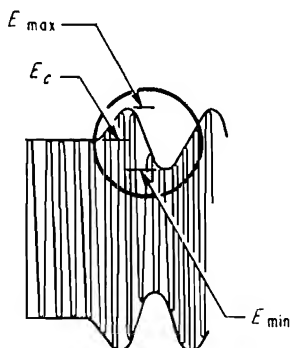


FIG. 13-29. Measurement of low levels of modulation.

two meters, one of which reads the average carrier level, and the other reads the percentage of modulation. The carrier-level meter will also indicate *carrier shift*. This term refers to the change in average carrier amplitude as modulation is changed from 0 to 100 per cent. These instruments are also provided with an overmodulation lamp which will flash when the modulation exceeds the value set on a dial.

A purely sinusoidally modulated signal of known modulation percentage is seldom available for calibration purposes. The instrument may, however, be calibrated by feeding in the output of two signal generators

TABLE 13-8. VALUES READ ON A VACUUM-TUBE VOLTMETER*

Ratio %	Average %	Positive peak %	Negative peak %
0		0	0
3.1	3.1	3.1	3.1
7.2	7.2	7.0	7.3
13.2	13.2	12.7	13.6
21.7	21.3	20.3	22.6
33.3	31.8	29.7	35.1
49.0	44.6	40.4	51.9
70.4	57.9	51.0	73.7
100	66.1	57.1	1

*See MacDonald (105).

which are separated by an audio frequency within the range of the instrument. The levels of the two signals can be accurately read with a vacuum-tube voltmeter. The indicated percentage of modulation will vary with the ratio between the signals. If the ratio between the smaller and larger signal is measured, the values that the meter should read are given in Table 13-8.

The reason for this variation is that the two beating signal generators produce a signal which is like a single-tone dsb suppressed carrier signal or a two-tone ssb suppressed carrier signal. When the calibration is being made, the carrier meter must always read 100, and both signal generators can be adjusted until this is so. This type of monitor is not at all suitable for reading ssb modulation because the modulation meter would read zero on a single tone. Single-sideband modulation is best checked with an oscilloscope. The maximum permissible peak amplitude is determined with a two-tone test. A normal ssb two-tone signal is shown in Fig. 13-30a while the overmodulated condition is shown in (b).

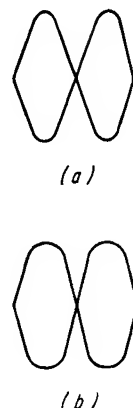


FIG. 13-30. Ssb envelopes for two-tone test. (a) Normal; (b) distorted.

Trapezoidal Pattern. It is easier to recognize a straight line than a perfect sine wave; for this reason a trapezoidal modulation pattern on an oscilloscope is useful. This type of pattern is obtained by connecting a sample of the modulated r-f output of the transmitter to the vertical plates (not through an amplifier) and a sample of the audio voltage at the modulator output to the horizontal plates. The sine-wave pattern as shown in Fig. 13-27 is obtained by using the internal oscillator sweep circuits (linear time base) which means that the spot moves at the selected speed from left to right but has almost instantaneous *flyback* from right to left. For the trapezoidal pattern the spot follows the audio in both horizontal directions.

In Fig. 13-31, showing various trapezoidal patterns, the unmodulated carrier is shown in part (a), and the pattern in part (b) is less than 100 per cent modulated. A correct 100 per cent modulated pattern is shown at Fig. 13-31c. Here the height of AB will be just twice AB in part (a). An overmodulated pattern is shown in part (d), and AB is more than twice the height of the unmodulated carrier. The pattern shown in part (e) occurs when there is insufficient r-f drive to the final amplifier, and the positive modulation is insufficient. If radio frequency gets on the horizontal plates, the pattern will lean over as shown in Fig. 13-31f.

Frequency Modulation. Frequency swing may be accurately measured by the "carrier disappearance" method. A number of ratios of fre-

quency swing to audio modulation were given in Sec. 7-3. The carrier may be observed on any receiver narrow enough to discriminate between the carrier and the first pair of sidebands. The most convenient type of receiver to use is a spectrum analyzer or a panoramic adapter. The

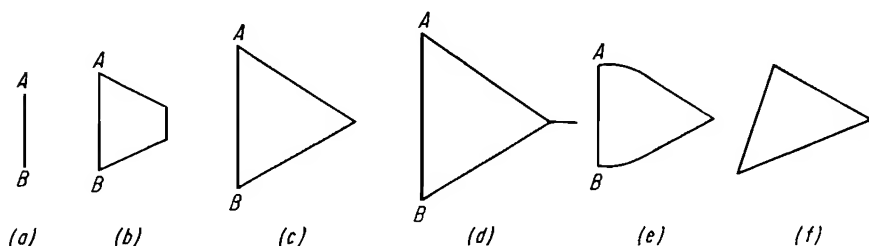


FIG. 13-31. Trapezoidal modulation patterns. (a) Unmodulated carrier; (b) less than 100% modulation; (c) 100% modulation; (d) overmodulation; (e) insufficient r-f drive; (f) r-f leakage to horizontal circuit in oscilloscope. (Courtesy of ARRL.)

audio frequencies required to give the frequency swing shown for the first and second disappearances are given in Table 13-9.

TABLE 13-9. AUDIO FREQUENCIES REQUIRED FOR SPECIFIC FREQUENCY SWINGS

Frequency swing, kc	First disappearance, cps	Second disappearance, cps
± 75	31,190	13,580
± 25	10,390	4,520
± 15	6,240	2,720
± 5	2,080	906

For example, suppose that it is desired to check the 100 per cent modulation mark on an f-m monitor for use in the 88- to 108-Mc band where the deviation is specified as 75 kc. An audio generator would be connected to the transmitter audio input terminals and set at 13,580 cps. A sample of the r-f output from the transmitter would be connected to the spectrum analyzer or mixed down to the range of a communications receiver. As the audio input is increased, the carrier will disappear when the frequency swing is ± 32.6 kc, come back up again, and disappear for the second time when the swing is exactly ± 75 kc.

F-M Monitors. As with a-m broadcasting, the FCC requires continuous monitoring of the modulation of an f-m broadcast transmitter. The monitor is also usually used for audio-distortion measurements, and the type of f-m discriminator used should preferably have not more than 0.25 per cent harmonic distortion. Monitors using tuned-circuit-type

discriminators have been made (106-108); but the usual commercial monitor uses a pulse-counter type of discriminator (109). A simplified circuit of the pulse-counter discriminator is shown in Fig. 13-32. As long as the capacitor loses practically all its charge through the resistor after each pulse, a meter in series with R will read a d-c value which is closely proportional to the number of pulses per second that are fed into C . The meter then becomes a "pulse counter." The 150-kc signal is removed by the low-pass filter, and the recovered audio will be an undistorted demodulated voltage, provided that the mixer and i-f amplifier stages pass all the sidebands of the modulated signal.

The difference between f-m and TV monitors is that 100 per cent modulation on an f-m broadcast is equivalent to ± 75 kc, while for TV it is equivalent to ± 25 kc.

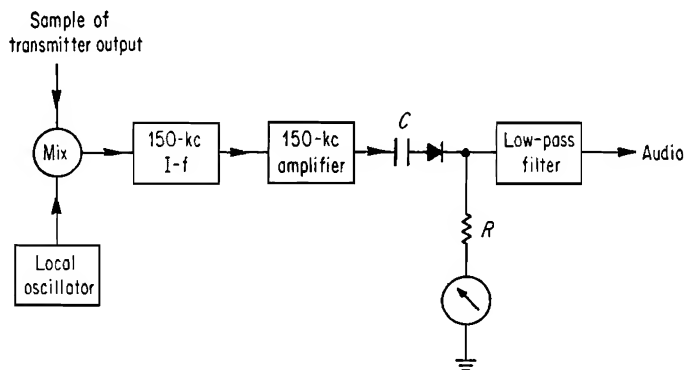


FIG. 13-32. Pulse counter discriminator.

In f-m broadcast and TV sound monitors, the 75- μ sec deemphasis circuit is used ahead of the audio terminals but must *not* be used ahead of the modulation indicator or overmodulation flasher. The modulation indicators will only show the correct deviation or swing at the high audio frequencies if the frequency response of the modulation meter is flat. The FCC permits the use of the deemphasis circuit in the audio output which is used for noise and distortion measurements.

Most monitors have two audio outputs. One is for connection to a monitoring speaker amplifier and is usually about 0 dbm across a 600-ohm balanced line. The other audio output is for noise, response, and distortion measurements and is at a level of about 10 volts across a high-impedance unbalanced (one side grounded) output. These are the expected levels for 100 per cent modulation at frequencies below 400 cps. At higher frequencies the level for 100 per cent modulation will drop according to the standard deemphasis curve.

An f-m transmitter, of course, cannot be amplitude-modulated to provide a reference level for the residual a-m measurement. A low-frequency reference level can be provided at the monitor or the rectified d-c at the a-m detector can be measured.

Other Methods of Measuring F-M. Carrier deviations produced by complex waveforms can be measured to an accuracy of 5 per cent and, with care, to an accuracy of 2 per cent when measuring sinusoidally modulated carriers, by means of the following method.

A sample of the transmitter output is mixed with the output of a signal generator in a crystal diode detector. The detector output is connected to the vertical-deflection terminals of a scope. The transmitter is modulated with the desired modulating signal, and this signal is also connected to the horizontal-deflection terminals of the scope. When the scope levels are properly adjusted, a rectangle will appear on the scope. When the signal-generator output frequency or one of its harmonics coincides with the transmitter frequency, a zero beat indication will be seen on the scope. This beat will be more easily seen if the modulating frequency is below about 300 cps. It will be found that the beat can be moved from left to right along the pattern by changing the signal generator frequency. The limits of the frequency excursion can be determined by measuring the signal generator frequency when the beat is at the left edge of the pattern and again when the beat is at the right edge of the pattern. One-half the difference between these two frequency measurements will be the frequency deviation. The left edge can be made the lower-frequency limit by selecting the correct polarity for the connections from the modulating-frequency generator to the transmitter. Instruments called transfer oscillators are available which have the scope, mixer and signal generator built in. The modulation is connected to the external horizontal sweep input, and the frequency is measured at the frequency-meter terminal by any of the methods discussed in Sec. 13-3. Once the voltage required at the transmitter input to obtain a certain deviation for one audio frequency is known, the deviation for any other audio frequency can be determined by measuring the frequency response. This method will not work with phase modulators.

The modulation can be recovered for distortion or response measurements by making use of an electronic frequency meter as a linear discriminator. The mixer output is connected to the frequency-meter input, and the signal generator is set for some value between 50 and 100 kc as read on the meter. The "pulse output" terminal on the frequency meter is then connected through a low-pass filter to a vacuum-tube voltmeter or scope, where the modulation can be observed. The low-pass filter must pass the desired modulation frequencies and stop the 50- to 100-kc signal. The low-pass filter should pass direct current so

that a calibration can be obtained. The direct voltage for a given unmodulated input frequency should be accurately measured. For example if the direct voltage at the output of the low-pass filter is 10 volts for 100-kc input and 9 volts for 90-kc input, ± 5 kc deviation will result in a 1-volt peak-to-peak or 0.353-volt rms output from the filter.

The term *frequency meter* may lead to some confusion. In one commercially available transfer oscillator, the terminal marked "frequency meter" is intended for connection to an instrument which is capable of measuring the frequency of the internal oscillator which may be between 100 and 220 Mc. The electronic frequency meter referred to above is an instrument which reads frequencies between 3 cps and 100 kc, and when used for f-m measurements will be connected to the terminal marked "mixer output" on the transfer oscillator and not the one marked "frequency meter."

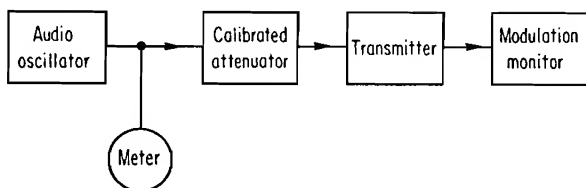


FIG. 13-33. Amplitude vs. frequency measurement.

13-6. Demodulated Audio Measurements. Audio-fidelity measurements are required by the FCC on a-m and f-m broadcast transmitters and on TV aural transmitters (110).

Frequency Response. Response measurements can be made by observing the variations on an input attenuator when a constant audio signal is fed into the attenuator and a constant modulation is maintained on the transmitter.

On an a-m transmitter 100 per cent modulation can be approached but never quite reached for these measurements; however, the modulation should be 100 per cent on an f-m transmitter. In fact, an f-m transmitter should be capable of reasonably good performance at 133 per cent modulation.

Response measurements can be conveniently made with a low-distortion audio-signal generator that has a built-in calibrated attenuator.

The preferred method is shown in Fig. 13-33. If a calibrated attenuator is not available, the method shown in Fig. 13-34 may be used. The resistor in each line must be half of the transmitter input impedance which may be either 600 or 150 ohms. It is important to notice that it is not fair to connect the meter directly across the transmitter input terminals. The reason for this is that the transmitter input impedance might not be the same value at all audio frequencies. If, for example,

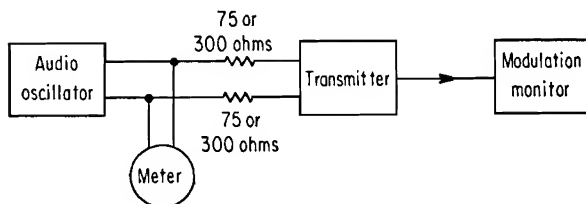


Fig. 13-34. Method of measuring audio input voltage.

the transmitter input impedance dropped from 600 to 200 ohms, the input voltage would drop to one-half, even though the meter reading in Fig. 13-34 were kept constant. This is what is supposed to happen during the test because, during actual operation, the voltage will be kept constant at the studio which will be at the far end of a line.

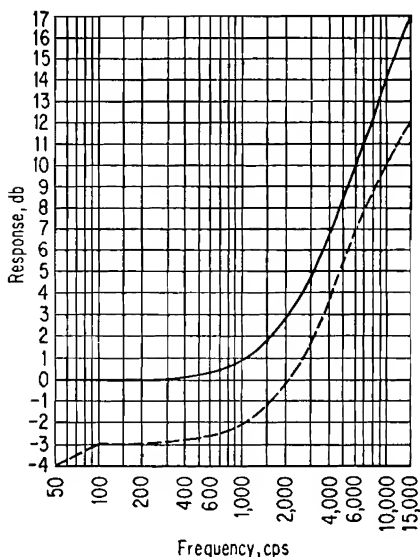


Fig. 13-35. Preferred response when 75 μ sec preemphasis is required.

Frequency response is generally expressed in decibels, and, when a voltmeter is used to record the response, the number of decibels for each frequency is given by

$$\text{Frequency response} = 20 \log \frac{E_{\text{ref}}}{E_f} \quad \text{db} \quad (13-26)$$

where E_{ref} = voltage required for the given modulation at reference frequency (1,000 cps for a-m and 400 cps for f-m)

E_f = voltage required to obtain same modulation at another frequency

A balanced H pad of 6 db or more may be used in place of the resistors.

The FCC allows a maximum departure of 2 db from the response at 1,000 cps for any frequency between 100 and 5,000 cps for a-m transmitters.

For f-m transmitters it should be possible to plot the response between the solid and dashed lines shown in Fig. 13-35. The solid line is the response given by a network having a time constant of 75 μ sec.

These limits refer to the complete system from microphone amplifier terminals to the antenna, and not more than half of this variation should be allowed in the trans-

Harmonic Distortion. Harmonic distortion is measured with single-tone modulation and is the ratio between the rms value of the single tone and the square root of the sum of the squares (rss) of all harmonics of the single tone. Typical distortion and noise meters are instruments which eliminate the fundamental tone with a tunable band-elimination filter. The residual distortion products and noise are then displayed on the meter.

The distortion products may be individually determined by means of wave analyzers. This type of measurement is not normally made at a transmitting station because it is not important to know the relative values of second, third, or higher harmonics.

When making distortion measurements, it is important that the audio-signal generator has low distortion. This should be measured directly before making measurements and should be less than 0.25 per cent.

Intermodulation Distortion. Another measure of the performance of any audio circuit is its *intermodulation* distortion. If a low frequency and a high frequency are simultaneously fed into an audio amplifier which has some distortion, additional frequencies which will be the sum and difference of the original frequencies will be generated. These are called the intermodulation (i-m) products. The i-m distortion is the ratio of the square root of the sum of the squares of the sum and difference frequencies to the amplitude of the higher frequency of the input signals times 100.

The usual practice is to make the low-frequency signal four times or 12 db higher than the high-frequency signal. Typical l-f values are 40, 60, and 100 cps, while 1,000, 7,000 and 12,000 cps are used for the higher frequencies. Under these conditions it is found that the i-m distortion is about 3.2 times higher than the harmonic distortion (h-d) for a single-ended audio amplifier and 3.8 times higher for a balanced push-pull amplifier (111-113).

Audio i-m tests are not normally used in routine transmitter testing because the test equipment required is more complex, and malfunctioning of a transmitter amplifier is revealed as much by the h-d test as by the i-m test.

The block diagram of a typical i-m test set is given in Fig. 13-36.

The commonest source of distortion in a transmitter is an amplifier which is being driven into a nonlinear region which may be either toward cutoff or toward saturation. This effect is illustrated in Fig. 13-37 where the dotted line shows the undistorted wave and the solid line shows the peak being compressed owing to some nonlinear condition. E_1 and E_2 refer to peak swings referred to the center frequency in an f-m transmitter or peak amplitudes referred to the carrier in an a-m transmitter. This type of distortion shows up as second harmonic when read on a

distortion meter or wave analyzer. The difference between E_1 and E_2 , expressed as a percentage of E_2 , will be (very approximately) equal to twice the harmonic distortion. This is not a satisfactory means of obtaining a distortion figure because of the difficulty of establishing the zero line with certainty and also because the distortion will usually be

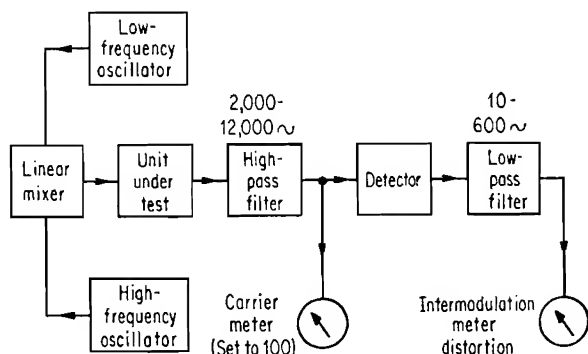


FIG. 13-36. Intermodulation distortion measurement.

beyond tolerable limits before being measurable by this method. A scope display of this kind, however, may be of assistance in clearing up distortion troubles in a transmitter.

In an a-m transmitter, flattening of the negative peaks may be due to an incorrect bias adjustment. Distortion of the positive peaks may be due to inadequate r-f drive or too high a plate load impedance (not enough antenna loading). In an f-m transmitter, mistuned r-f circuits can introduce distortion, but the trouble will almost always be in the modulator.

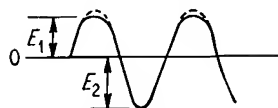


FIG. 13-37. Typical second harmonic distortion due to nonlinearity in transmitter amplifiers.

It is sometimes useful to make final transmitter adjustments while observing a distortion indicator; in fact, some transmitters actually have built-in distortion meters for this purpose. This is particularly valuable when adjusting bias, loading, and r-f drive to a class B linear amplifier (114) or when making intermediate-cavity adjustments in a klystron amplifier.

Noise Measurements. The measurement of noise output from a transmitter in the absence of modulation is simple in theory. A reference level on a distortion and noise meter is first established. The reference level is 100 per cent modulation as observed on a modulation meter and may be any audio frequency between 100 and 5,000 cps for an a-m transmitter but must be 400 cps or less for an f-m transmitter. The signal is then removed, the input is terminated with a resistance equal to the

rated source impedance (150 or 600 ohms), and the rms voltage is read again. The ratio in decibels is the residual noise.

Two complications are the nature of the noise and the characteristics of the measuring meter (115). The FCC specifies that noise measurements be made with an instrument having the "ballistic characteristics similar to those of a standard VU meter" (116).

The principal noise components in most transmitters are multiples of the power-line frequency, and in this case the meter characteristics will have little effect on the observation.

In the case of television stations, the visual transmitter may be switched off during noise measurements on the aural transmitter.

When residual amplitude modulation is being measured on an f-m transmitter, the reference level is obtained by measuring the d-c voltage at the output of the detector. The rms value of the noise is measured at the output of a 75- μ sec deemphasis circuit following the detector. The residual a-m noise is then given by

$$\text{Noise} = 20 \log \frac{0.707 E_{d-c}}{E_{rms}} \quad \text{db} \quad (13-27)$$

The FCC requires that noise measurements on an f-m transmitter be made only in the 50- to 15,000-cps band and that distortion products above 30,000 cps need not be observed. In the case of a-m transmitters noise measurements may be made in the 50- to 7,500-cps band, and distortion products up to 16,000 cps must be observed. This means that it is permissible to use low-pass filters ahead of the measuring equipment, but there is little advantage in the case of f-m transmitters because the 75- μ sec deemphasis circuit attenuates rapidly above 15,000 cps. In the case of a-m transmitters the FCC has an additional requirement that no objectionable interference be caused in the band outside the range of $\pm 7,500$ cps from the carrier frequency. Transmission above 7,500 cps should therefore be kept to a minimum.

Measurement Precautions. It is important that pessimistic hum and noise readings be avoided by paying proper attention to shielding and grounding. If the input of the equipment under test is balanced to ground, the audio generator should have balanced output terminals. If an attenuator is used in a balanced line from the generator to the equipment under test, there must be resistance balance on the two sides of the attenuator, and the capacity to ground on each side should be balanced. Similarly, if the input is unbalanced, an unbalanced attenuator should be used.

All high-impedance connections, such as the leads from the modulation monitor to the noise and distortion analyzer, should be made with

shielded wire. Such connections should not be too long because the cable capacity can cause false readings at the higher audio frequencies.

Detailed measurement procedures are given in a number of EIA and IRE Standards (117-121).

13-7. Telegraph Keying Measurements. Telegraph keying distortion in either on-off or frequency-shift keying is the departure of the wave-shape of the demodulated carrier from the desired waveshape (122). The desired waveshape is a compromise between the square wave for maximum intelligibility and waveforms which produce minimum adjacent-channel interference. Telegraph keying was discussed in detail in Secs. 6-4 and 12-3.

There are two types of keying distortion.

1. *Bias* telegraph distortion is that in which all mark pulses are lengthened (positive bias) or shortened (negative bias). It may be measured with a steady stream of "unbiased reversals," square waves having equal-length mark and space pulses. The average lengthening or shortening gives true bias distortion only if other types of distortion are negligible.

2. *Characteristic* distortion occurs in a circuit where a mark-to-space transition can start before the previous space-to-mark transition was completed. The effect of characteristic distortion therefore depends on the length of the transmitted pulses. This is a typical "long-line" effect and does not normally occur in a transmitter unless there are long time constants in the keying circuit.

A third type called *fortuitous* distortion is the result of interference to the circuit and is important in system tests but is not applicable to the transmitter itself.

A telegraph transmitter should be tested by an oscilloscope observation of keying reversals up to the highest operating speed. In addition, widely spaced minimum-length marking characters and widely spaced minimum-length spacing characters should be checked. The marking or spacing bias can be measured by adjusting the scope sweep rate to 1.5 times the keying speed. In this way the marking signals will overlay the spacing signals, and bias can be estimated to within 5 per cent. Another method is to measure the time interval of the keying signal and the recovered keyed signal on an oscilloscope with an accurately calibrated time base.

The keyed signal should also be examined for leading- or trailing-edge transients, amplitude variations, and breaks due to relay chatter.

Printer Margin Measurements. When machine printer signals are being sent, a well-adjusted printer may be used for measurement of telegraph distortion. The *telegraph range finder* on the machine is used to find the *margin* when a perfect test signal is fed into the transmitter.

On perfect test signals the range finder can be moved from 10 to 90 on the 0-to-120 scale before errors will start to occur. One hundred divisions on this scale corresponds to one unit of the printer code. The range finder effectively shifts the time of operation of the start-pulse-sensing device relative to the message-pulse-sensing devices. (See Fig. 13-38 which shows range-finder operation. This illustration also shows various types of distortion.)

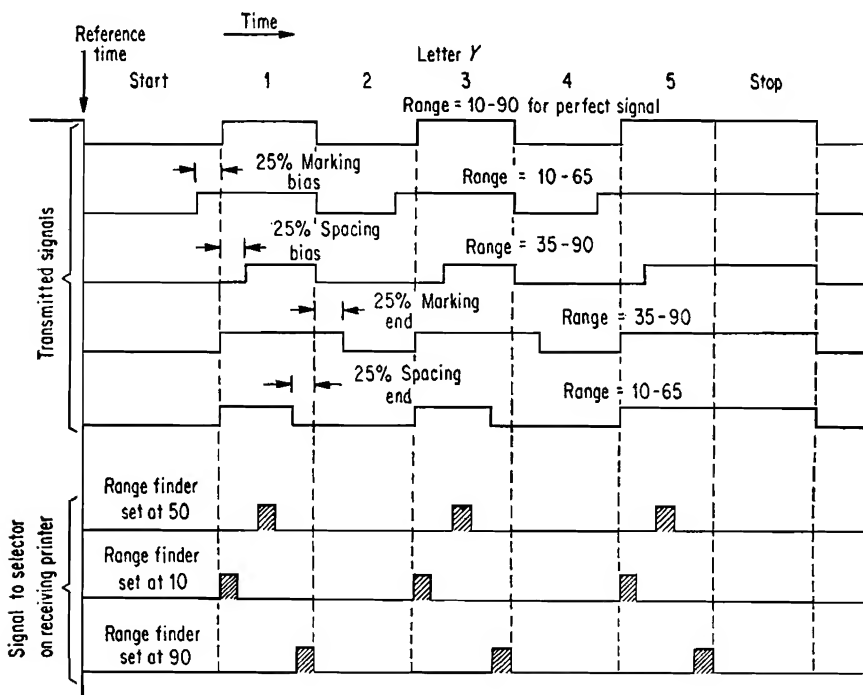


FIG. 13-38. Teletype distortion measurements.

Accurate distortion measurements can be made with special instruments. One type of instrument uses a stroboscopic method, and other types use a cathode-ray type of display. Synchronous signals are displayed on a circular trace on the scope, while a spiral trace is used for start-stop signals.

13-8. Broadband Radio System Measurements. In a broadband radio system such as one carrying 120 voice channels, an important measurement is the intermodulation between channels. This can be determined by the *noise-loading* test (123). In this method random noise is spread across the entire frequency spectrum with the exception of the channel selected for measurement. Noise is eliminated from this channel by means of a band-elimination filter. This noise signal is then

used to modulate the transmitter. The signal is then detected, and a measurement is made in the selected channel by means of a band-pass filter. The noise in this 3 kc "slot" will then be the total effect of the residual noise and the intermodulation products.

In another version of this test the noise is passed through a 50- to 500-kc band-pass filter for 120-channel circuits and through a 50- to 1,100-kc band-pass filter for 240-channel circuits (124). The intermodulation noise is then read in a 3-kc slot which is outside of the range in which the noise is being transmitted.

The objective for the nationwide toll dialing system in the United States is the maintenance of average noise below 31 dba for more than 50 per cent of the time and below 52 dba for more than 99.9 per cent of the time (125). (See Table A-5 in the Appendix for definition of dba.)

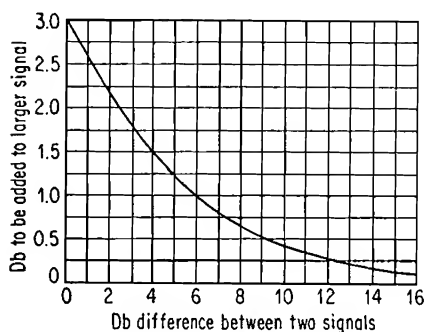


FIG. 13-39. Resultant noise increase when two sources of noise are combined. (Courtesy of Lenkurt Electric Co.)

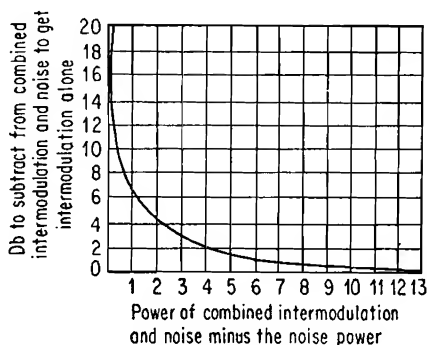


FIG. 13-40. Resultant noise reduction when two sources of noise are separated. (Courtesy of Lenkurt Electric Co.)

Measurements are made at a point in the circuit where the transmission level is 0 dbm. It is necessary to establish this reference level in telephone transmission measurement work because there is nothing equivalent to 100 per cent modulation to use as a reference. Typical broadband transmitter noise for one 3-kc channel is 14 to 18 dba. The intermodulation noise and all other noise introduced into the system must not bring the noise level above 31 dba for more than 50 per cent of the time.

Figure 13-39 shows how the noise is increased when two sources of noise are combined. For example, if the noise in a channel is 24 dba, and this is fed through some equipment which has a noise level of 14 dba, the difference is 10 db, and, according to Fig. 13-39, 0.4 db must be added to 24 dba, and the resulting level will be 24.4 dba.

Figure 13-40 shows how the intermodulation noise alone can be found if the transmitter noise alone and the combined transmitter noise and

intermodulation noise are measured. For example, if the total noise is 20 dba and this drops to 16 when the noise generator is switched off, the difference is 4 db. From Fig. 13-40, 2 db must be subtracted from 20, which gives 18 dba for the intermodulation level. This checks with Fig. 13-38, which indicates that 2 db must be added to the larger signal when there is a 2-db difference between the signals being combined.

13-9. Pulse-transmitter Modulation Measurements. Noise, response, and distortion measurements on a pulse transmitter will be made in essentially the same manner as with f-m or a-m transmitters. Appropriate demodulation equipment must be used, and the level corresponding to 100 per cent modulation must be established. An oscilloscope with a calibrated time base or an auxiliary timing generator must be used to check pulse widths. Methods of pulse measurement are given in an IRE Standard (126).

13-10. Monochrome Video Modulation Measurements. There are three general ways of checking the performance of a visual transmitter. These are

1. Quantitative measurement of power, frequency response, and hum level
2. Quantitative and qualitative check of the horizontal waveform (line) and vertical waveform (field) on an A scope (linear time base)
3. Qualitative check of transmitted picture on a picture monitor

Frequency Response. The frequency response of a TV transmitter could be determined in the same manner as is done with an a-m or f-m transmitter. That is, the clamps could be disabled and a fixed bias applied to the modulated amplifier; then the input voltage required to produce a given modulation could be determined. The resulting data could then be compared with the "ideal detector curve," and this would satisfy one FCC requirement. However, this method would not reveal the response of the lower sideband and, since no sync pulses are present, the transmitter would not be driven to its peak output. Moreover, the FCC specifies that "a composite signal composed of a synchronizing signal to establish peak output voltage plus a variable-frequency sine-wave voltage occupying the interval between synchronizing pulses" shall be applied for measuring attenuation characteristics.

In Fig. 13-41a is shown an ideal monochrome transmitter output. If this signal is detected with a wideband diode detector, the a-f response will be as shown in Fig. 13-41b. Now, if the transmitter response between -0.75 Mc and $+0.75$ Mc is as shown by the dotted line, the diode detector output would remain unchanged. It is clear, therefore that some form of narrow-band measuring instrument is necessary to check

the lower sideband. This could be done with a spectrum analyzer but is more usually done with an instrument called a *sideband analyzer* (127).

Sideband Analyzer. The sideband analyzer generates a video signal that sweeps from 0 to 7 Mc and includes a receiver whose local oscillator is swept in synchronism with the video. A sweeping voltage is also available to permit the horizontal deflection of a scope to be synchronized with the sweep. The instrument also includes a calibrated notching marker. Tuned circuits in the receiver must be set up for the particular channel being used.

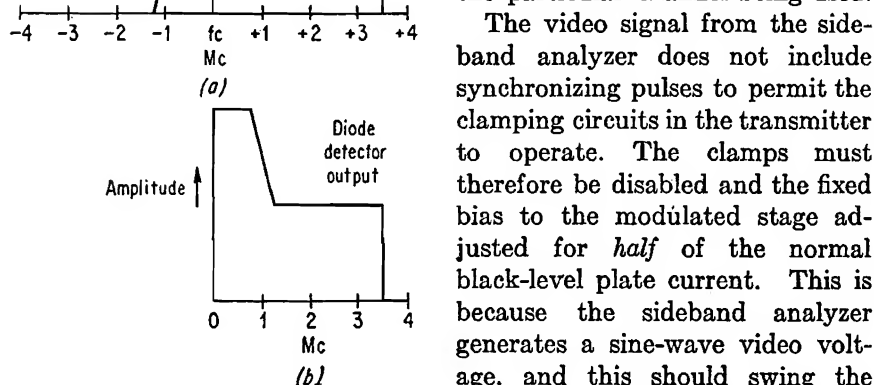


FIG. 13-41. Ideal monochrome television transmitter outputs. (a) Sideband output; (b) ideal detector output.

circuit so that a base line can be established on the scope. The video voltage is therefore only applied to the transmitter for half of the time.

To use the instrument, practically any type of oscilloscope can be connected to the *V* and *H* terminals of the cable provided with the unit. A sample of the output voltage should be picked up from a directional coupler which is on the antenna side of the sideband filter, diplexer, and harmonic filter. The video sweep output voltage should be connected through all filters, equalizers, and distribution amplifiers that are normally used but should not go through a stabilizing amplifier. The detector peak control on the analyzer should be adjusted for maximum signal on the scope. Then the video output control should be adjusted so that the sidebands are just one-half the amplitude of the carrier. The correct sideband analyzer display is shown in Fig. 13-42. A transmitter having this response would be satisfactory for color transmission. For monochrome, no response above 3.5 Mc is required. The FCC requires

that all visual transmitters have at least 20-db attenuation at 4.75 Mc and above.

When the instrument is connected as described above, and the display of Fig. 13-42 is used, the transmitter may be adjusted and the sideband filter (if one is used) may be checked. This arrangement, however, does not meet the requirements of Sec. 3.687 of the FCC Rules and Regulations Covering Radio Broadcast Services because of the omission of the synchronizing pulses.

The sideband analyzer may be used in conjunction with a video-transmission set to provide a composite signal and an independent check of the upper and lower sidebands. The output from a test oscillator should be connected through the video-transmission set so that an accurate point-by-point check may be made. The equipment arrangement is shown in Fig. 13-43. In this case the amplitudes may be adjusted by observation of the waveform scope. The sync should be adjusted to

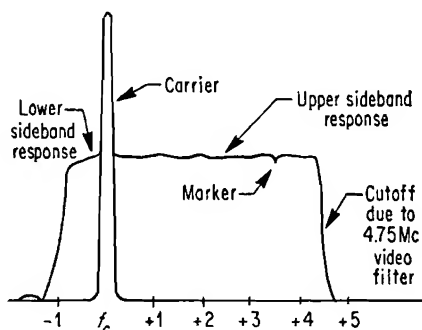


FIG. 13-42. Sideband analyzer display. (Courtesy of RCA.)

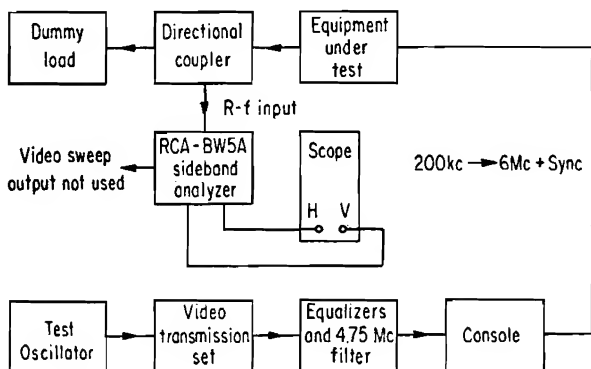


FIG. 13-43. Measurement of visual transmitter frequency response.

the proper peak and the proper video level should be established by feeding in a signal at some frequency between 200 and 500 kc. The video level should be adjusted for full black-level to white-level swing as shown on the waveform scope, and the voltage required to do this should be read at the output of the test oscillator. This voltage should then be used for all test frequencies and the resulting amplitude read off the scope. Typical amplitudes for 0.7 and 1 Mc are shown in Fig. 13-44.

For this test, of course, the clamps are restored in the transmitter, and it will be operating in a completely normal manner.

The low-amplitude levels which will occur at frequencies more than 1.25 Mc below carrier can be read by calibrating the scope with a known voltage and then increasing the scope gain so that the lower sidebands can be seen. The scope calibration is repeated at the increased gain and the ratio between lower-sideband voltages and the voltage at 200 kc can then be determined.

The "ideal detector response" which is required by the FCC is determined by merely adding the upper- and lower-sideband voltages together at each frequency.

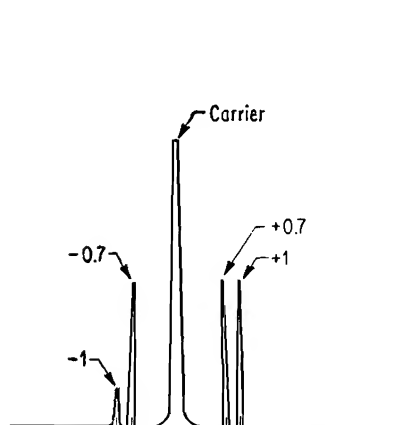


FIG. 13-44. Sideband analyzer display for single video frequencies.

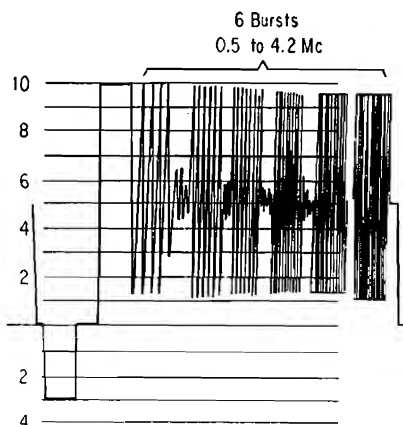


FIG. 13-45. Waveform display of multi-burst signal.

Multifrequency Burst Check. A quick system check of frequency response is available by using a *multifrequency burst* signal. Such a signal is available in some transmission test-signal generators. This signal consists of six "bursts" of different test frequencies that are sent in sequence during the horizontal interval. The amplitudes are viewed on a waveform scope. The amplitudes should remain constant through all video amplifiers but at frequencies above 1.25 Mc should drop to 50 per cent of the low-frequency amplitude when viewed with a wideband detector at the transmitter output. The amplitudes should be restored to a constant value when the signal is viewed on a properly adjusted demodulator that attenuates the lower frequencies by 6 db. A typical multiburst signal is shown in Fig. 13-45. The white reference signal is used for measuring frequencies below 150 kc and can also be used as a reference for other purposes.

Sweeping Oscillators. The frequency response of a television transmitter or any broadband circuit may be checked by means of a sweeping

oscillator. The oscillator may be swept slowly with a motor, making use of a scope with a long-persistence screen. Many electronically swept oscillators are available. The usual sweep rate is 60 cps. In order to calibrate the resulting display on a scope, it is necessary to insert a "marker." In some oscillators the markers are crystal-controlled signals and provide a positive sharp pulse.

Most r-f circuits being checked will show some variation in input impedance over the frequency range of interest. It is very important that the sweeping oscillator output remain constant at a point which is on the generator side of a resistance equal to the load impedance. This is illustrated in Fig. 13-46 and was mentioned under audio tests for a-m and f-m transmitters. Thévenin's theorem states that any complex network of generators and impedances can be replaced by a single constant-voltage generator and a single series resistance which is equal to the impedance looking back into the network. Any generator that automatically maintains a constant voltage at the input of the series resistor regardless of frequency or load may be termed a *Thévenin generator*. This is much more important in r-f work than in video work because there is not much difficulty in making the input impedance of a video modulator or distribution amplifier look very close to 75 ohms.

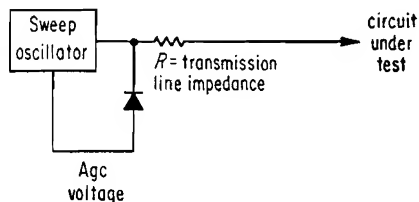


FIG. 13-46. Thévenin generator.

A good Thévenin generator can be made with a motor-driven oscillator which obtains its plate voltage from an amplitude-regulating power supply. The advantage of the motor drive in preference to the much faster electronic sweep is that the power supply has a chance to respond to abrupt changes in load impedance and to maintain a constant voltage at the input to the 50-ohm resistor even though the voltage at the other side may be changing violently. The equipment arrangement for this type of sweeper is shown in Fig. 13-47.

There is available a broadband generator which can be set on center frequencies between 200 kc and 1,000 Mc and can be swept with a sweep-width variable from 100 kc to 300 Mc (128). This is also a Thévenin generator as it has a full-wave detector circuit ahead of an internal 50-ohm resistor. The detector controls an automatic-gain-control (agc) circuit.

Direct Point-by-point Measurements. A more tedious method of frequency response checking is to merely use a signal generator on the input and a suitable indicator at the output, as is done in audio checking. At microwave frequencies, a vacuum-tube voltmeter cannot be conveniently

used, and the usual method is to use a crystal diode detector as shown in Fig. 13-48. The signal generator is modulated at 1,000 cps, and an amplifier which is sensitive only to a 1,000-cps signal is used. A sensitive d-c meter could be used in conjunction with an unmodulated signal from the generator, but it is much more difficult to avoid unwanted hum and noise pickup than with the modulated method. Variable-frequency tuned amplifiers are also available.

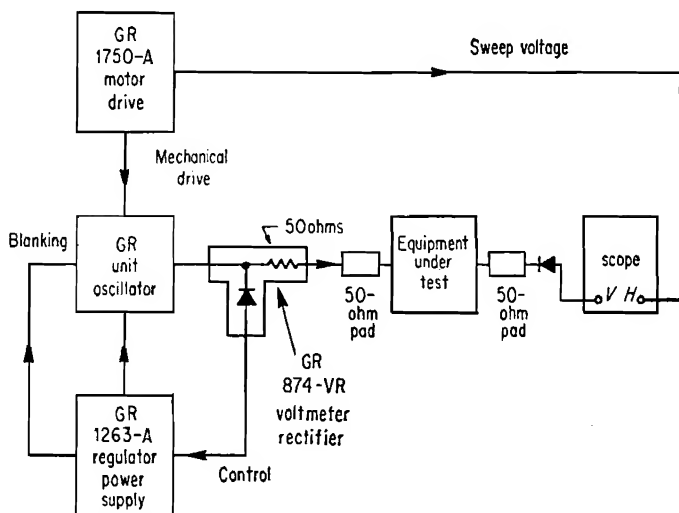


FIG. 13-47. Motor-driven sweep oscillator connections. (Courtesy of General Radio Co.)

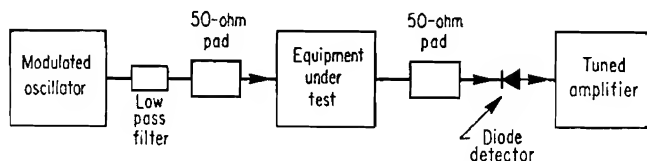


FIG. 13-48. Use of tuned amplifier for microwave measurements. (Courtesy of Hewlett-Packard Co.)

It is important to note that the diode detector in combination with a scope or meter cannot discriminate between fundamental and harmonics. It is good practice, therefore, to include a harmonic filter after the signal generator, which passes the frequencies of interest but eliminates second and higher harmonics. It is also good practice to isolate the equipment under test from both the generator and the detector with 50-ohm attenuation pads having an attenuation of at least 6 db.

An alternate type of indicator which operates with the signal generator unmodulated and gives a good indication with low output signals is the calibrated i-f amplifier. A typical i-f amplifier operates at 30 Mc and

has a dynamic range of 80 db. It is necessary to heterodyne the output signal down to 30 Mc with a separate oscillator and mixer as shown in Fig. 13-49.

Circuit Response at Low and High Signal Levels. It is very important to note that the response of any nonlinear circuit may not be the same at signal generator levels as at operating levels. For example, the input impedance of an r-f amplifier will change as it is driven into the grid-current region. For this reason, the response of a transmitter amplifier may be quite different at signal generator and operating levels. This is why it is desirable to adjust a television transmitter with an instrument such as the sideband analyzer.

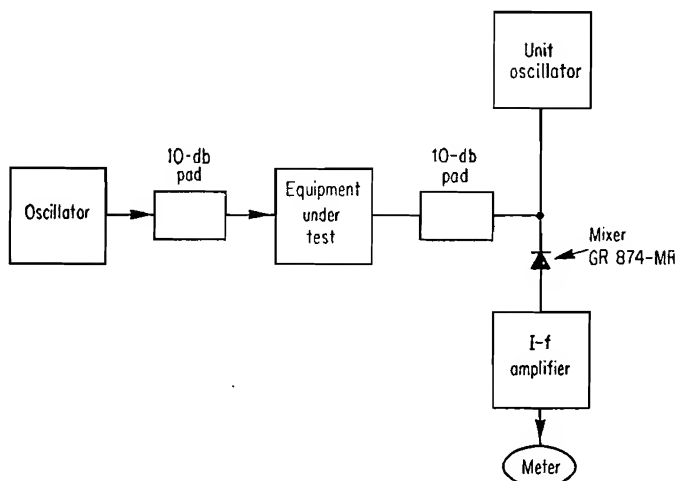


FIG. 13-49. Use of i-f amplifier for microwave measurements. (Courtesy of General Radio Co.)

Klystron amplifiers also exhibit some change in frequency response as the drive level is increased.

Any circuit which contains no nonlinear elements such as a vestigial-sideband filter or harmonic filter will have the same frequency response at signal generator and operating levels. Of course, this is true only if there is not sufficient heating to cause any detuning effects.

Measurements Using Pulse or Transient Techniques. The effect of a visual transmitter on a picture which arrives at its input terminals depends not only on the frequency response but also on the phase response. Methods of checking the phase response will be discussed under color measurements. If a square wave is applied to the input terminals of a visual transmitter, the detected wave at the output will depend on both amplitude and phase responses. The question is frequently asked: "Why not use square waves for testing video circuits?"

While it is true that equipment which is performing correctly will have a certain square-wave response, it is not easy to identify the trouble if the response is *not* satisfactory. For this reason the FCC does not specify a square-wave response, and the method is not generally used when testing installed transmitters. A complete discussion and bibliography on video measurements employing transient techniques has

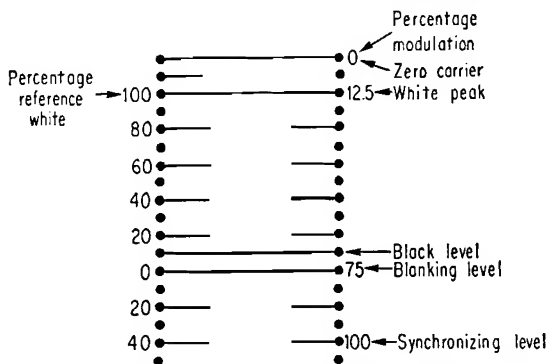


FIG. 13-50. Waveform scope operating scale. (Courtesy of RCA.)

been given by Samulon (129). The shape of the desired square-wave response was discussed in Sec. 12-9. Methods of measurement are given in an IRE Standard (130).

There are some advantages to using a testing signal which has its frequency spectrum limited (131). A *sine-squared* pulse is such a signal. A typical sine-squared pulse generator provides three sine-squared pulse widths: 0.25, 0.125, and 0.062 μsec , corresponding to bandwidths of 2, 4, and 8 Mc, respectively.

Waveform Display. One of the best checks on the performance of a TV transmitter is the waveform display. It is the practice in the

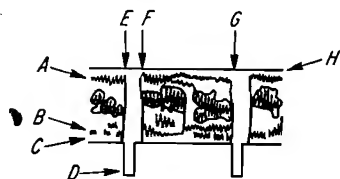


FIG. 13-51. Monochrome horizontal scanning interval. (Copyright 1965 ATT Long Lines Dept.)

United States to display the waveform signal "upside down." We usually expect the oscilloscope trace to move upward with increased output, and this is the type of display always used with the sideband analyzer. The maximum output from a visual transmitter occurs at the *peak of the sync*, and this is always shown at the bottom of a waveform display while a line representing zero carrier output is at the top of the display. A standard operating scale for use on a waveform scope is shown in Fig. 13-50 (132).

It is usual practice to sweep the waveform scope at one-half the line rate (7,875 cps) for the line display. This results in two sync pulses per

line as shown in Fig. 13-51. Signal measurement terms have been standardized by the IRE (133). The various intervals and levels shown in Fig. 13-51 are as follows:

- A: white peak
- B: black peak (Black level is the maximum excursion of the black peaks)
- C: blanking level (The setup is $CB/CA \times 100\%$)
- D: sync level
- E-G: horizontal interval = $63.5 \mu\text{sec}$
- E-F: horizontal blanking interval = $10.05 \mu\text{sec}$
- F-G: picture signal = $53.45 \mu\text{sec}$
- H: zero carrier line

The zero carrier line is only available at monitoring points where a d-c voltage exists that can be "chopped." Such a d-c voltage will be available at diode detector outputs but not at the outputs of video amplifiers. The common method is to use a relay which has its contacts connected to short out the detected d-c and coil connected to a source of 60 cps. Vacuum-tube or transistor switches can also be used.

One basic waveform check is that the *polarity* of the input signal to the transmitter is correct. In audio work this does not matter, but the visual transmitter output must increase as the signal goes toward black. It is standard in the United States for the input signal to the transmitter to be *black negative*. The signal will be inverted as it goes through each vacuum-tube stage in a video amplifier, and it is necessary for all amplifiers in the signal path to have an even number of stages so that both input and output will always be black negative. It is also standard to make crystal diode detectors so that the d-c output voltage will be negative, thus giving a black negative waveform. The standard input voltage to a transmitter is 1.4 volts peak to peak as measured from the sync peak to the reference white level. This is convenient because the left-hand 0-to-100 scale in Fig. 13-50 will correspond to 1 volt peak to peak.

An expanded portion of the waveform signal near the horizontal blanking interval is shown in Fig. 13-52. The various intervals and levels are as follows:

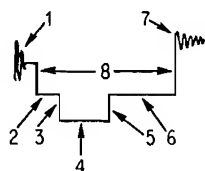


FIG. 13-52. Monochrome horizontal blanking interval. (Copyright 1955 ATT Long Lines Dept.)

- 1. Video at right side of picture
- 2. Front porch $1.59 \mu\text{sec}$
- 3. Leading edge of horizontal sync pulse
- 4. Tip of sync $4.76 \mu\text{sec}$

5. Trailing edge of sync
6. Back porch 4.76 μsec
7. Video at left side of picture
8. Horizontal blanking interval 11.11 μsec

The objective in adjusting transmitter modulation levels is to make the sync peak amplitude 40 per cent of the blanking level to white peak amplitude. To make sure that receivers which use intercarrier sound detection always have enough level from the visual transmitter to generate a satisfactory 4.5-Mc signal, the FCC requires that the reference white level shall be 12.5 ± 2.5 per cent of the peak carrier level. In Fig. 13-51, the space between A and H must be 12.5 per cent of the amplitude from D to H. The FCC requires that the setup shall be 7.5 ± 2.5 per cent of the blanking level to white peak level, or C to B must be 7.5 per cent of C to A.

If the waveform scope is swept at 30 cps, the vertical-scanning interval can be examined. This is shown in Fig. 13-53. The peak levels will be the same as in Fig. 13-51. This display will reveal 60- or 120-cps hum

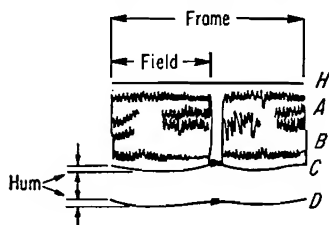


FIG. 13-53. Monochrome vertical scanning interval. (Copyright 1955 ATT Long Lines Dept.)

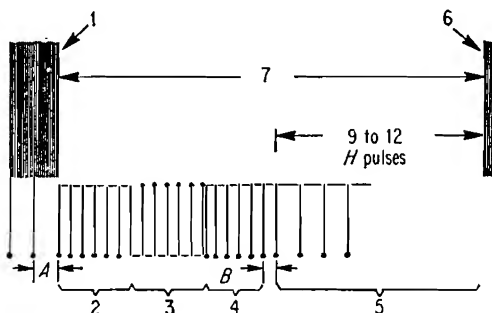


FIG. 13-54. Monochrome vertical blanking interval. (Copyright 1955 ATT Long Lines Dept.)

or other low-frequency variations. The FCC specifies that the peak-to-peak (p-p) variation within one frame shall not exceed 5 per cent of the peak signal as measured at both the sync peak and blanking levels. In Fig. 13-53 the p-p changes at C and D must not be more than 5 per cent of D to H.

An expanded portion of the waveform signal near the vertical blanking interval is shown in Fig. 13-54. The pulse groups are as follows:

1. Video at bottom of picture
2. First six equalizing pulses
3. Serrated vertical sync pulse
4. Second six equalizing pulses
5. Nine to twelve horizontal pulses with video blanked to allow receiver scope to retrace to top of picture—one less pulse on alternate lines
6. Video at top of picture
7. Vertical blanking interval

Because of the interlaced method of scanning, intervals *A* and *B* will be interchanged on alternate fields.

Trouble Indications on the Waveform Display. There are many types of transmitter trouble that will show up on a waveform display, but only a few will be mentioned. The time intervals shown in Fig. 13-52 are not usually changed by a transmitter and depend only on the performance of the synchronizing generator. The transmitter may distort waveforms and analysis of the transmitter performance is done by comparing the signal appearing at various points in the transmitter with the input signal.

Sync Compression. Most transmitters will compress the sync signal somewhat and compensation is achieved by "stretching" the sync in the modulator. If compression is excessive, this is an indication that some r-f amplifier is not delivering the required peak power, and an adjustment of r-f drive or loading may be needed.

White Compression. If the modulator drives the transmitter to full peak output before driving the white level down to within 12.5 per cent of zero carrier, the bias on the modulated amplifier may be too high. Another cause is that the r-f driver may have poor regulation and delivers more output as the modulated amplifier is driven towards cutoff. This is corrected by adding resistive loading or "swamping" to the r-f driver.

Clamp Failure. Failure of a clamp circuit will show up in the vertical blanking interval. Referring to Fig. 13-54, the blanking level lines will slope down to the right. Unbalance in a keyed clamp circuit will appear as an uneven back porch.

Waveform Oscilloscope Characteristics. The IRE has published standards on television measurements which specify the response characteristics of the oscilloscope used for waveform measurements (134). This is done so that negligible errors will be introduced by waveform overshoots showing up differently on different scopes. The response characteristic is called *IRE roll-off* and is shown in Fig. 13-55. This response can be obtained by using the circuit shown in Fig. 13-56. The purpose of this standard is somewhat similar to the reason behind setting up the vu standard for audio measurements. When many people are observing the same signal at different points in a system with different instruments,

it is important that the instruments have similar characteristics. The IRE roll-off response *must not* be used when measuring color burst or chrominance amplitudes.

Picture Monitor. The quality of transmitted pictures is also checked with a picture monitor. These instruments usually include both a waveform scope and a picture tube (135). Impairment of the transmitted picture is a subjective quantity and is measurable only in terms of the

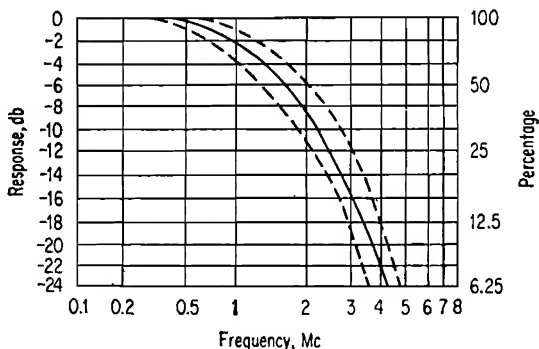


FIG. 13-55. IRE roll-off response characteristic. (Courtesy of IRE.)

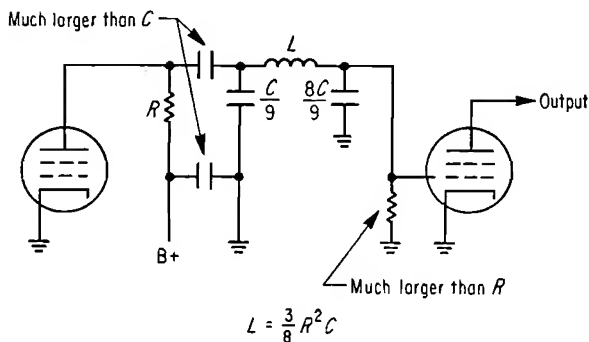


FIG. 13-56. Relative circuit values to produce IRE roll-off. (Courtesy of IRE.)

reaction of the observer (136). To assist in the judgment of the picture, standard resolution charts have been prepared. These charts include a gray scale and a resolution wedge. A well-adjusted camera can resolve 600 lines horizontally across the picture, but, after the signal has been passed through a transmitter followed by a vestigial-sideband filter, the resolution will seldom exceed 350 lines.

Picture monitors are often provided with a switch that gives a display called the *pulse cross* (137,138). This is a display of the lines occurring in the neighborhood of the vertical-retrace period. The triggering of the

scope is displaced both vertically and horizontally so that the shadows produced by the sync and blanking pulses appear near the center of the picture tube. The pulse cross is more useful in checking a synchronizing generator than a transmitter. A pulse cross is shown in Fig. 13-57.

Picture Monitor Pickup. It is very important to keep in mind the fact that the *receiver-attenuation* (RA) method is used in television (139). This means that receivers must be 6 db less sensitive at the carrier frequency than at frequencies above 1.25 Mc. This in turn means that a transmitter performance must *not* be judged by checking the picture

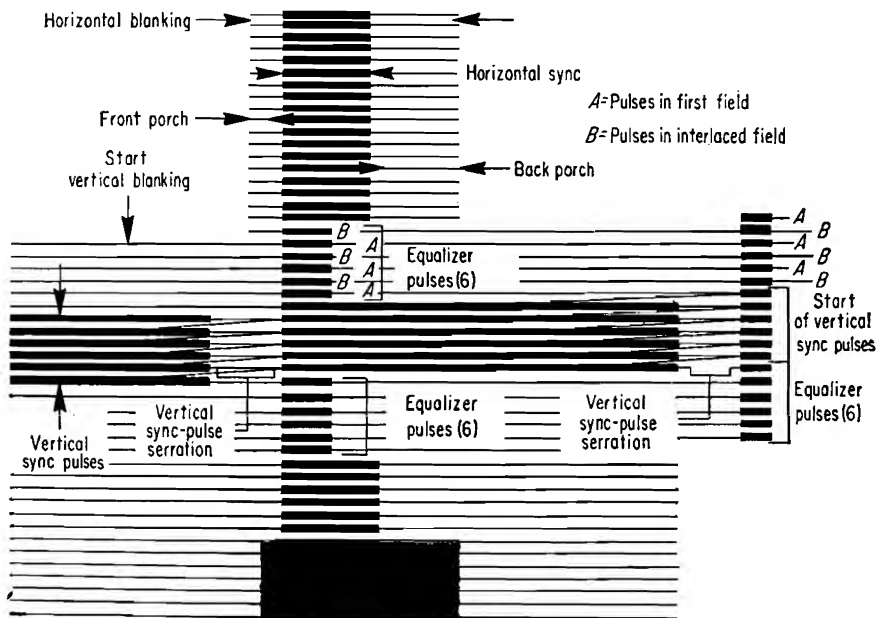


FIG. 13-57. Expanded pulse cross display. (Courtesy of Electronics.)

with a wideband detector (such as a crystal diode detector directly on a directional coupler) at a point where the lower sideband has been attenuated. Most r-f amplifiers including the modulated stage will start to cut off the lower sideband. If the picture is checked at the output of any r-f amplifier, the high-frequency response will be unfairly attenuated. Figure 13-58 shows an ideal transmitter output response, ideal detector output, and an ideal receiver response. The detector which is used for checking the picture at the transmitter output or any r-f amplifier output must have a response as close as possible to the ideal receiver response, and this should be maintained in a stable manner. This can be done with a stable transmission-line-type filter or a crystal-controlled superheterodyne receiver. A typical receiver has a sound notch filter so that

it can be used for monitoring with the aural transmitter on. For taking measurements with the aural transmitter off, the notch filter can be switched out to give improved high-frequency phase and amplitude response.

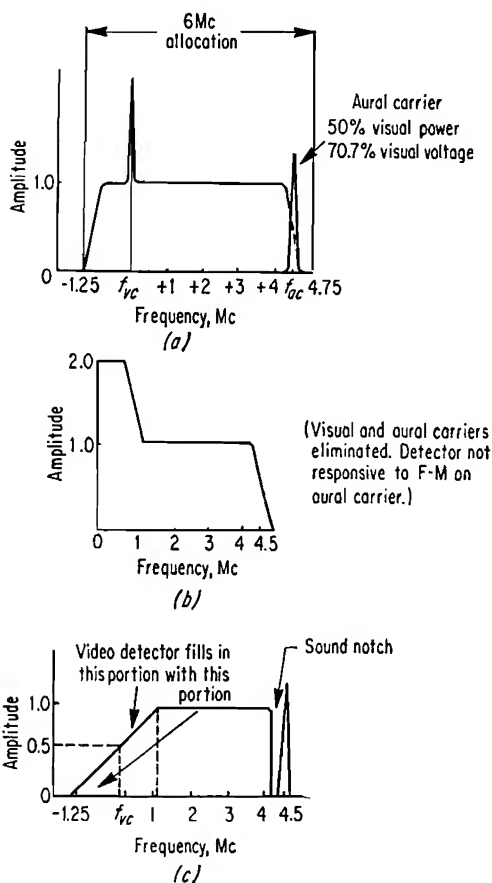


FIG. 13-58. Transmitter and receiver responses for RA television transmission method. (a) Transmitter output; (b) output of diode detector with no preceding filter; (c) ideal receiver i-f response.

13-11. Color Video Measurements. Additional transmitter measurements are required to ensure that there will be satisfactory transmission of a standard color signal. The FCC specifies additional frequency response, a measurement of envelope delay and characteristics of the color signal. Additional measurements, not specified by the FCC, such as differential phase response and differential amplitude response, are frequently made.

Envelope Delay. Envelope delay was defined and discussed in Chaps. 3 and 12. Phase-correcting filters are required ahead of the video input to the transmitter for three principal reasons:

1. Compensation for phase characteristics of typical receivers.
2. Compensation for filters following the transmitter such as vestigial-sideband filter, harmonic filter, or diplexer.

3. Compensation for other video filters such as 4.75-Mc cutoff filter. Correction must be applied so that the envelope delay specified by the FCC is met. This is shown in Fig. 13-59. The desired objective is that, in the over-all system including the transmitter and the receiver, all frequencies in the band of interest should travel through the system

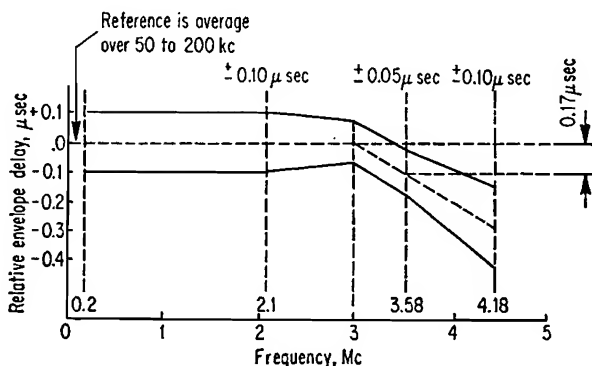


FIG. 13-59. Envelope delay limits specified by FCC for color transmitters.

in the same length of time. If they do not, there will be poor registration of the colors over the luminance signal. This is sometimes called the "funny-paper effect."

A quick check of the amplitude and phase response of a system can be made with the sine² pulse (140). The camera signal resulting from a sharp black-to-white transition is very close to a sine² pulse. A square wave which has edges shaped like those of a sine² pulse permits the low-frequency attenuation and phase characteristics of a system to be checked. The high-frequency response of the system will not have a critical effect on these observations. A typical generator provides a composite signal consisting of a horizontal sync pulse, a sine² pulse and a square wave with sine² edges. Quantitative comparisons with previous data can be made at the points indicated in Fig. 13-60. Low-frequency response is indicated by tilt of the square-wave flattop, and high-frequency response is indicated by height of the sine² pulse relative to the height of the square wave and also by overshoot on the leading and trailing edges of the square wave. The phase characteristic is indicated by symmetry of the sine² pulse about its vertical axis.

It is important to perform this test with a demodulator having phase and amplitude characteristics which will not degrade the signal. The effect of a longer transmission time for the higher frequencies is to produce ringing on the demodulated pulse which is greater on the trailing than on the leading edge. The opposite effect occurs if the transmission time for the higher frequencies is shorter. As the pulse width of the input signal is narrowed, the output signal amplitude will drop, and the ringing on the leading and trailing edges will increase. The typical test pulse has a pulse width at the 50 per cent amplitude point which is equal to *twice* the period of the highest frequency to be transmitted. This is

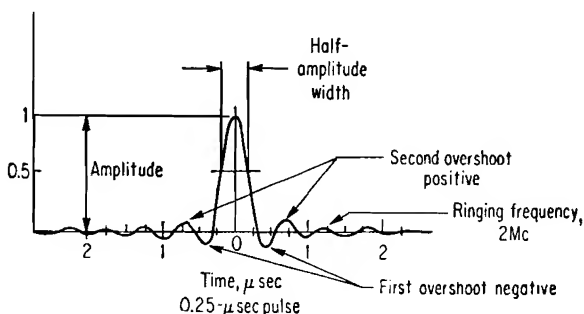


FIG. 13-60. Sine-squared pulse display. (Courtesy of Telechrome Mfg. Corp.)

called the $2T$ pulse. For a typical TV transmitter this is the $0.25\text{-}\mu\text{sec}$ pulse. The commonly used pulse widths and corresponding bandwidths at the half-amplitude point are shown in Table 13-10.

TABLE 13-10

Pulse width, μsec	6-db Bandwidth, Mc
0.250	2.0
0.125	4.0
0.062	8.0

For a transmitter giving good performance the pulse width of the demodulated pulse will not exceed $0.375\text{ }\mu\text{sec}$ for the $0.25\text{-}\mu\text{sec}$ input pulse. The leading and trailing first overshoots will be less than 5 per cent of the pulse amplitude. Component performance will vary, and a *rating factor* has been used to identify the performance of individual components.

These methods do not give the quantitative measurements required to demonstrate compliance with FCC rules. If two frequencies which are 200 kc apart are passed through the transmitter and the change in the relative phase between the two frequencies is measured, envelope delay is given by

$$T_e = \frac{\phi_2 - \phi_1}{2\pi(f_2 - f_1)} \quad \mu\text{sec} \quad (13-28)$$

where T_e = envelope delay, μsec

ϕ = phase-angle shift as signal goes through system, deg

f = frequency of each signal, Mc

This will give one value of envelope delay at the average frequency between the two selected. A series of values is then taken up to about 4.3 Mc. Note that it does not matter what the absolute value of the envelope delay is, and it is the *change* in envelope delay that must not exceed certain limits.

In one equipment which has been made for numerical checks, the phase angles are displayed on a meter (141). In other equipments a sweeping technique is used, and the two frequencies are applied as subcarriers on top of the video-modulating frequency (142). The demodulated signal appears as a trace on a scope. These instruments can be directly calibrated by inserting a piece of delay line with known delay between the test-signal output terminals and the demodulator.

Color-signal Characteristics. The characteristic of the color signal is checked by feeding standard color signals through the equipment and checking the demodulated signal on a scope which displays amplitude and phase (143,144). Standard color signals can be generated by special instruments for this purpose (145).

The transmitter output is first fed to an r-f demodulator, and the video output is then connected to a color demodulator. The demodulated colors can then be displayed as a spot on a cathode-ray tube which has a face scale calibrated directly in amplitude and phase. This display was discussed in Chap. 12 and is given in Fig. 12-12. The FCC specifies that the angles must be within 10° and the amplitudes must be within 20 per cent of the correct values at 75 per cent of full amplitude. The various color signals must also be checked with the waveform display, and the ratios of the subcarrier amplitudes to the luminance signal amplitudes for each color must be between 0.8 and 1.2 of the correct values for 75 per cent of full amplitude. The correct values are shown in Fig. 12-11.

Amplitude Linearity. The FCC specifies that the *transfer characteristic*, which is the relationship between the transmitter r-f output and the video input signal, shall be substantially linear between reference black and reference white levels. The color check described above also indirectly specifies amplitude linearity but no numerical limits for linearity are given.

Linearity can be conveniently checked with a *stairstep* generator (146-150). If a color subcarrier signal is superimposed on each step, the amplitude of the subcarrier at each step in the output signal gives a measure of *differential gain*. If the detected r-f signal is passed through a high-pass filter, the "steps" are eliminated, and the amplitudes of the

subcarrier at each step can be conveniently compared. A modulated stairstep signal can be generated with typical video-transmission test-signal generators. A high-low cross filter can be used to eliminate the steps to compare the subcarrier modulation or to eliminate the subcarrier

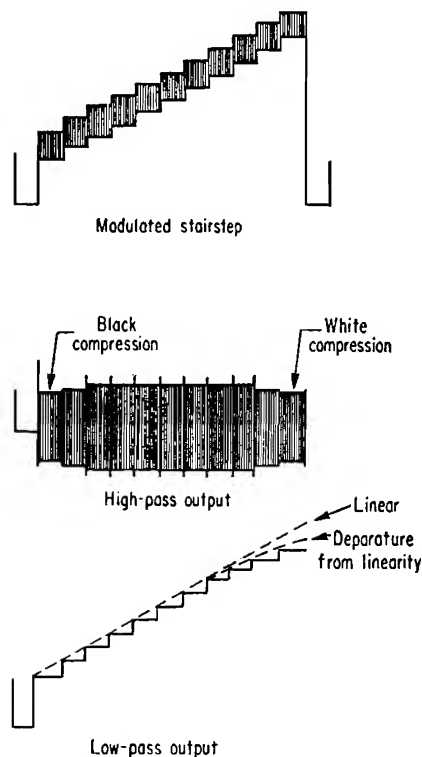


FIG. 13-61. Differential gain measurement with stairstep generator. (Courtesy of Telechrome Mfg. Corp.)

modulation and leave the steps to check the black-to-white linearity. Figure 13-61 shows a modulated stairstep and outputs from the high and low positions of the cross filter.

Differential Phase. Another important measurement is the change in phase shift of a given video frequency as the amplitude is changed. The important video frequency in a color-television system is 3.58 Mc. Instruments are available which permit this measurement to be made with an accuracy of 0.25° . A phase-reference signal is generated during the horizontal sync

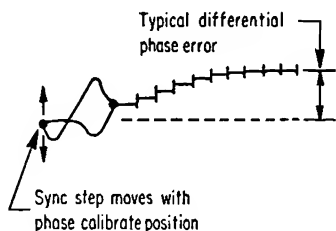


FIG. 13-62. Display for differential phase measurement. (Courtesy of Telechrome Mfg. Corp.)

pulse, and the phase angle at each step of a stairstep signal is converted to an amplitude display on a scope. The height of the line on the scope is related to the phase, and the measurement is made by matching the reference signal with each step being checked. The reference line is moved by means of a front-panel calibrated control on a typical instrument. A typical display is shown in Fig. 13-62. The maximum phase variation should not exceed $\pm 5^\circ$.

Vertical-interval Test Signals. All of the tests mentioned so far are necessarily made during nonbroadcasting hours. A useful test signal can be sent during the last part of the vertical interval, and this can be sent at any time during broadcasting (151-157).

The FCC permits a test signal to be sent which commences with the last 12 μ sec of the 17th line and extends through the 20th line after the start of vertical blanking. A guard interval of at least $\frac{1}{2}$ -line duration at blanking level must separate the test signal from the start of the picture. The picture should commence not later than line 22. Modulation of the television transmitter by these test signals must be confined to the area between the reference white level and blanking level except where a chrominance subcarrier is being sent in which case the negative excursion may extend into the synchronizing peak amplitude. Test signals are not permitted during the horizontal blanking interval. The test signal may be a multiburst, stairstep, color bar, or any other desired signal. If the test signal includes a peak reference white signal, this signal may be used for automatic-level control purposes (158).

13-12. Antenna Measurements. All of the measurements which have been made up to this point can be carried out with the transmitter connected to a dummy load. Before connecting the transmitter to the antenna, some check should be made to ensure that power will be delivered to the antenna. Most high-powered vacuum tubes and klystrons will be quickly destroyed if they are operated in an unloaded amplifier.

Only those antenna measurements that concern a transmitter operator will be discussed in this chapter. In the case of f-m or TV omnidirectional antennas, microwave reflectors, rhombic antennas, and curtain arrays there is nothing that the transmitter operator can do to change the directivity, and it is only necessary for him to satisfy himself that the impedance "looking in" to the antenna feeder can be handled by the transmitter. On the other hand, a directional antenna operating in the 540- to 1,600-kc broadcast band has a directivity which is very much under the control of the station operator.

Impedance Measurements with Bridges. At frequencies below 60 Mc, antenna impedances can be measured directly with radio-frequency bridges. Typical instruments of this type read resistance directly at any frequency in the range but the reactance scale readings must be multiplied by a frequency correction factor. A signal generator and a well-shielded receiver are needed to use these instruments.

At frequencies above 30 Mc, antennas are almost always connected directly to a transmission line without adjustable tuning elements and measurements are usually made at the end of the line rather than directly at the antenna terminals. A vhf bridge is available which is suitable for direct measurements below 200 ohms in the range of 10 to 165 Mc. Another type of bridge samples the electric and magnetic fields in a transmission line. This instrument gives readings directly in terms of impedance and phase angle rather than separating out resistance and reactance. The phase-angle scale is direct-reading at 100 Mc but must be corrected

at other frequencies. The observed phase angle is multiplied by the frequency in megacycles and divided by 100. The resistance and reactance can be obtained from the impedance and phase angle as follows

$$R = \frac{Z}{\sqrt{1 + \tan^2 \theta}} \quad (13-29)$$

where Z = impedance, ohms

θ = angle, deg

R = resistance, ohms

After R is determined, the reactance is obtained from

$$X = R \tan \theta \quad (13-30)$$

Other types of bridges have been described for impedance measurements (159-161).

One type of bridge has a self-contained signal generator and a null detector (162).

Measurements with Admittance Meter. A null device is available which balances in the same manner as a bridge. This is a very simple instrument with five sexless coaxial connectors and three adjustable loops. Connections are made to a generator, detector, standard resistance, a one-eighth-wavelength stub (or capacitor), and the unknown to be measured. The three loops are adjusted for a null, and the susceptance, conductance, and multiplying factor are read directly from a calibrated scale. A simplified schematic of the instrument is shown in Fig. 13-63, and the scale is illustrated in Fig. 13-64.

This instrument does not need to be corrected for frequencies between 41 and 1,000 Mc. The range can be extended down to 20 Mc and up to 1,500 Mc by applying a correction factor to the susceptance reading. The range for both conductance and susceptance is 0.2 to 1,000 mmhos. When a one-quarter-wavelength line is connected between the instrument and the unknown, multiplication of the scales by 2.5 will give resistance and reactance directly over a range from 1 to 5,000 ohms. The instrument will also read reflection coefficient and impedance magnitude in terms of voltage ratios observed with specific loop positions.

Impedance Variation along Transmission Lines. If the antenna is not exactly matched to the transmission-line feeder, there will be a standing wave on the line. The voltage-standing-wave ratio (vswr) is a measure of the degree of mismatch and becomes very large if the line is opened or shorted at the end. If there is a standing wave on the line and the impedance is measured where the voltage is at a maximum, the impedance will be a maximum and will be purely resistive. Similarly, the impedance will be a minimum and will again be purely resistive at the voltage mini-

imum point. At other points on the line the impedance will include a reactive component.

The output coupling circuit of a transmitter may be unable to handle the maximum or minimum impedance on a line driving a mismatched

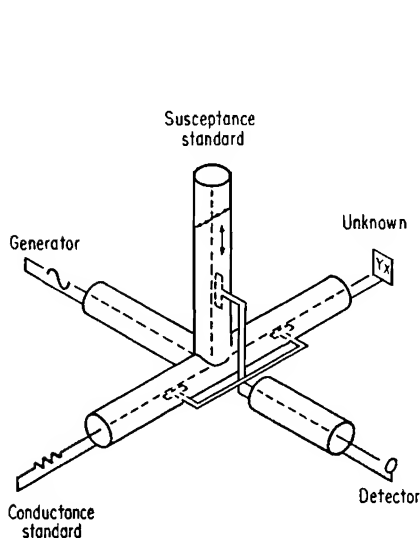


FIG. 13-63. Simplified schematic of admittance meter. (Courtesy of General Radio Co.)

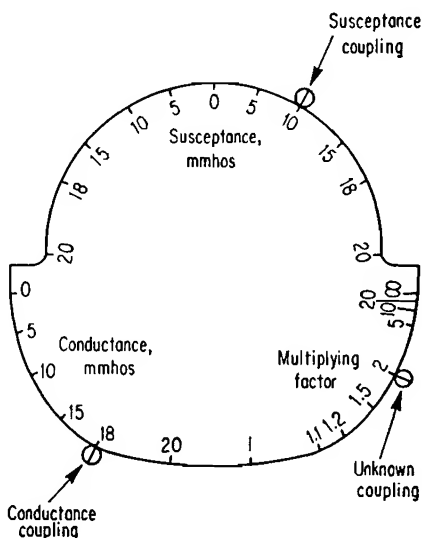


FIG. 13-64. Admittance meter scale. (Courtesy of General Radio Co.)

antenna but may be able to handle an intermediate value of impedance even if some reactance is present. For this reason, a small change in the length of transmission line between the transmitter and antenna may permit more power to be delivered to the antenna. The length should not be changed more than one-half wavelength.

TABLE 13-11. TYPICAL V_p VALUES

Type of dielectric	Velocity of propagation V_p (as fraction of speed of light)
Solid polyethylene.....	0.66
Solid teflon.....	0.70
Semisolid polyethylene.....	0.84
Air.....	Over 0.99

One-half wavelength on the line will be *shorter* than in free space by a factor which depends on the *velocity of propagation* (163). This factor has typical values shown in Table 13-11, but exact values should be obtained from the cable manufacturer.

One-half wavelength of line will be given by

$$\frac{\lambda}{2} = \frac{V_p 492}{f} \text{ ft} = \frac{V_p 5,905}{f} \text{ in.} \quad (13-31)$$

where V_p = velocity of propagation

f = frequency, Mc

λ = wavelength in desired units

When it is desired to measure the impedance of an antenna or other component that cannot be conveniently connected directly to a bridge, an approximate measurement can be obtained by inserting any exact number of half wavelengths of line between the bridge and the unknown. The length of line chosen can be checked by placing a known impedance at the end and checking that the bridge shows this value.

Impedance Measurements with Slotted Lines. Slotted transmission lines can be used for measurements of impedance at frequencies where at least one-half wavelength will appear on the line. The impedance can be determined from a measurement of the voltage-standing-wave ratio and the distance from the load to a voltage maximum or minimum. The impedance must be calculated or determined from a *Smith chart* (164). Slotted lines are not often used in transmitting stations because the transmitter operator usually does not need to know the exact impedance of a vhf or uhf antenna. It is sufficient to determine the voltage-standing-wave ratio alone, and this does not require a slotted line.

Slotted lines are provided with scales calibrated in millimeters and a tunable detector probe. The usual practice is to feed about 1 mw into the line from a signal generator modulated at 1,000 cps and to read the signal with a tuned amplifier. A 6- to 10-db attenuator should always be used to isolate the line from the generator.

Smith Chart. A Smith chart is shown in Fig. 13-65. This chart has a special impedance coordinate system which portrays the impedance at any point along a transmission line in relation to the impedance at any other point. The scales around the outside of the chart are calibrated in fractions of a wavelength, and the full circle (360°) corresponds to one-half wavelength. Resistances and reactances are plotted as fractions of the impedance of the transmission line. The resistance lines are circles which are tangent to the bottom of the chart, and the reactance lines are portions of circles which are tangent to the vertical line through the chart which is called the *axis of reals*. The chart has many uses including general impedance transformations, stub-tuner design, and finding input impedance when the load and length are known (165-167).

An example of determination of the load when the voltage-standing-wave ratio, line impedance, and distance to a voltage minimum are known is shown in Fig. 13-65. Instead of measuring from the voltage

minimum all the way to the load it is convenient to merely short the load and measure the new position of the voltage minimum.

In the example shown, which is for a lossless line, the voltage standing-wave ratio was measured as 3.7. The reflection coefficient circle is drawn with center at $R = 1$ and $X = 0$ and with the circumference

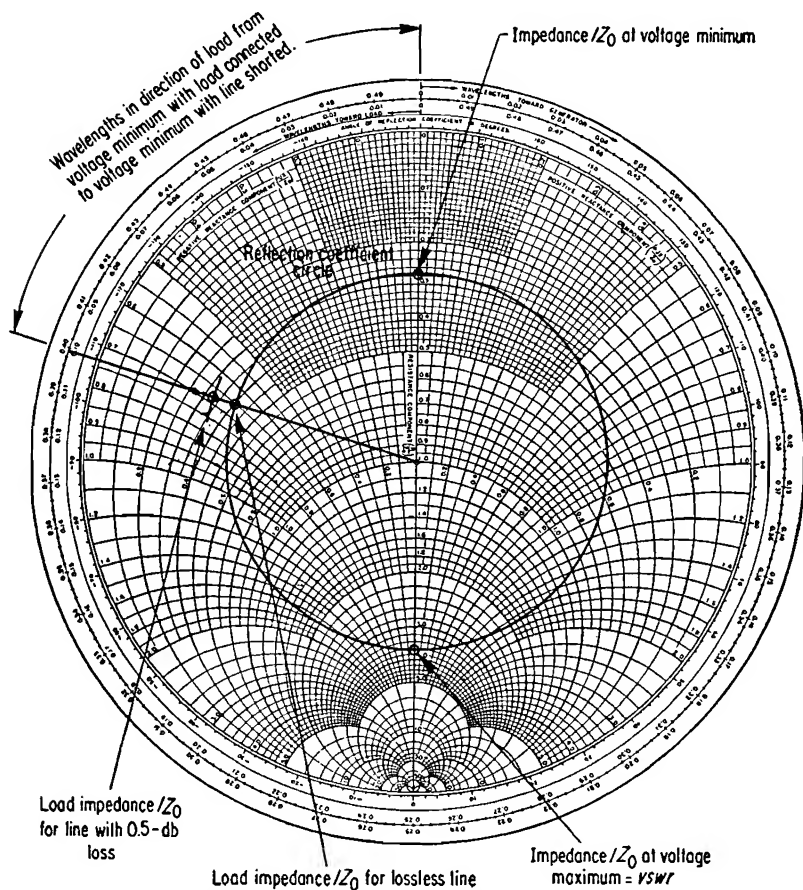


FIG. 13-65. Smith chart. (From Philip H. Smith, *Transmission-line Calculator*, *Electronics*, January, 1939, and January, 1944.)

cutting the axis of reals at 3.7. With the load connected, a voltage minimum appeared at 62.7 mm, and with the load shorted the minimum moved to 79.6 mm. With the load still shorted, the next minimum towards the load was found to be at 159.6 mm. One-half wavelength on the line was then $159.6 - 79.6 = 80.0$ mm. This could also have been determined by converting the frequency of the signal generator to wavelength and multiplying by the velocity of propagation on the slotted line,

The full wavelength was 160 mm, and the fraction of a wavelength from the "load" minimum to the "short" minimum was

$$\frac{79.6 - 62.7}{160} = 0.1057$$

A line drawn through the circumference at 0.1057 to the center of the circle cuts the circle at a point which gives the load impedance. The slotted line had been set up with the zero end of the scale toward the generator. When the load was shorted, the minimum moved toward the load so the "wavelengths toward load" direction was used to find the 0.1057. The intersection point is $0.41 - j0.65$. The slotted line had an impedance of 50 ohms, and so the actual load impedance was $50(0.41 - j0.65) = 20.5 - j32.5$ ohms.

Now consider the case where the line between the slotted line and the load has some attenuation. When the line has attenuation, the voltage-standing-wave ratio "improves" as the observer moves towards the generator.

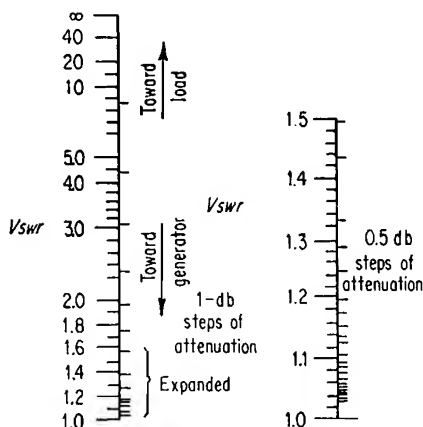


FIG. 13-66. Effect of line attenuation on voltage-standing-wave ratio. (From "Reference Data for Radio Engineers," IRT.)

ratio to about 4.5. This will move the reflection-coefficient circle out, and the new intersection point will be about $0.33 - j0.68$ ohms.

There are sources of error in slotted-line measurements, but these are small and will not be discussed here (168-170).

An instrument is available which displays the impedance by means of a spot of light on a translucent Smith chart. Other types of charts can be used to present impedance in polar coordinates.

Directional-coupler Measurements. The transmitter operator can usually obtain all the necessary information about the antenna from a

generator. The improvement ratio constantly changes as the voltage-standing-wave ratio changes; the relationship is shown in Fig. 13-66. No numbers are shown on the decibel scales because the zero must be taken at the point where the voltage-standing-wave ratio is read in. Suppose that in the above example the cable between the slotted line and load had an attenuation of 1 db per 100 ft and the cable length was 50 ft. The total attenuation would then be 0.5 db. The voltage-standing-wave ratio was 3.7, and it is seen from Fig. 13-66 that 0.5 db towards the load brings the voltage-standing-wave

directional coupler. Most vhf and uhf transmitters will have built-in directional couplers and these will be connected to meters which read the forward or *incident* power or the reverse or reflected power. The ratio between these two powers in decibels can be converted to a voltage-standing-wave ratio from Fig. 13-67. In the case of f-m transmitters, a low

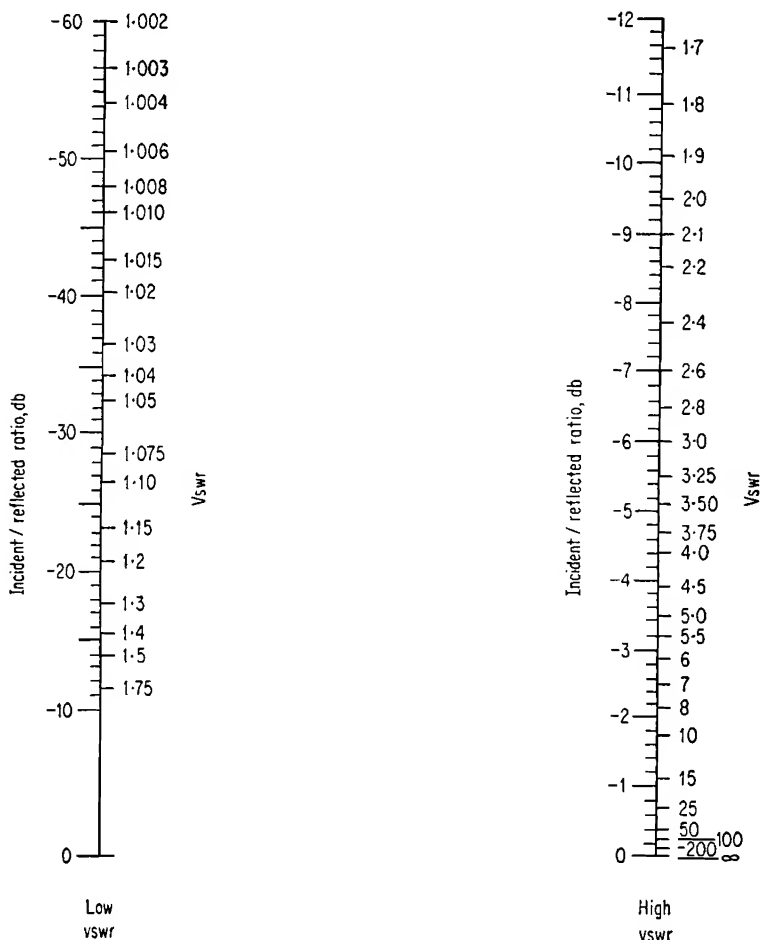


FIG. 13-67. Relation between voltage-standing-wave ratio and ratio between incident and reflected power. (From "Reference Data for Radio Engineers," ITT.)

voltage-standing-wave ratio is important only from an efficiency consideration; but in the case of television transmission the voltage-standing-wave ratio has an important effect on the picture. If the antenna causes a bad reflection, the signal may be reflected again from the transmitter to be retransmitted as a "ghost" on the picture. A television antenna

and its associated transmission line should have a voltage-standing-wave ratio of 1.1 or better.

Location of Transmission-line Faults. In case a transmission line or waveguide develops a poor contact at a joint, it is important to determine the location of the trouble before sending a man up the tower to try to repair it.

Two basic methods of fault location are the *pulse-echo* method and the *swept-frequency* method. The pulse-echo method is more applicable to long telephone cables and will not be discussed in detail (171,172). In brief, the time taken for a pulse to travel to the discontinuity and back is measured. The distance to the fault is then

$$D = 492V_p t \quad (13-32)$$

where D = distance to fault, ft

V_p = velocity of propagation as percentage of velocity of light

t = total time for pulse to go and return, μsec

If the input frequency (c-w) to a transmission line is varied, two variables can be measured to determine fault locations. These are

1. Voltage-standing-wave ratio at the line input
2. Impedance at the line input

The voltage-standing-wave ratio will not vary with frequency unless there are two faults on the line. It is sometimes convenient to introduce a mismatch at a known location on the line in order to determine the position of an unknown fault. For example, a 25- or 75-ohm resistor could be placed at the end of a 50-ohm line. The voltage-standing-wave ratio would then vary in a cyclic manner as the frequency is changed and the distance between the points of mismatch would be given by

$$D = \frac{492V_p}{f_2 - f_1} \quad (13-33)$$

where D = distance between faults, ft

V_p = velocity of propagation of cable as percentage of velocity of light

f_1 = frequency at point of maximum voltage-standing-wave ratio, Mc

f_2 = frequency of next highest point of maximum voltage-standing-wave ratio, Mc

In the case of waveguide the fault separation is

$$D = \frac{492}{f_2 - f_1} \sqrt{1 - \left(\frac{f_c}{f}\right)^2} \quad (13-34)$$

where D = distance between faults, ft

f = transmitter operating frequency, Mc

f_c = cutoff frequency of waveguide, Mc

f_1 = frequency at a point of maximum voltage-standing-wave ratio, Mc

f_2 = frequency at next highest point of maximum voltage-standing-wave ratio, Mc

The cutoff frequency of waveguide is always less than the operating frequency and is given by

$$f_c = \frac{590}{\text{WR number}} \quad (13-35)$$

where f_c = cutoff frequency, kMc

WR number = long dimension of waveguide cross section, hundredths of in.

For example the cutoff frequency of WR-975 is 590/975 or 0.605 kMc or 605 Mc.

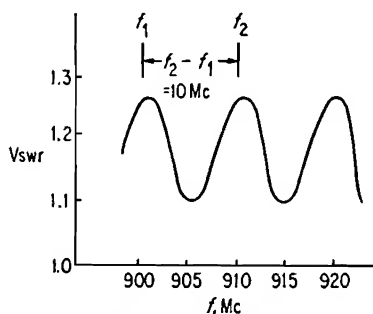


FIG. 13-68. Typical variation of voltage-standing-wave ratio with frequency for two faults on line.

A typical plot of voltage-standing-wave ratio against frequency for two faults on a line is shown in Fig. 13-68. If the voltage-standing-wave ratio due to each fault in the absence of the other is represented by r_1 and r_2 , then

$$r_1 = \sqrt{r_{\max} r_{\min}} \quad (13-36)$$

$$r_2 = \sqrt{\frac{r_{\max}}{r_{\min}}} \quad \text{for } r_1 \text{ greater than } r_2 \quad (13-37)$$

and
$$r_{\max} = r_1 r_2 \quad (13-38)$$

$$r_{\min} = \frac{r_1}{r_2} \quad \text{for } r_1 \text{ greater than } r_2 \quad (13-39)$$

where r_{\max} and r_{\min} are the maximum and minimum values of voltage-standing-wave ratio as frequency is changed.

When the faults are quite close together, the frequency difference between vswr maxima may be many megacycles apart, and other frequency-sensitive discontinuities may show up which will mask the observations.

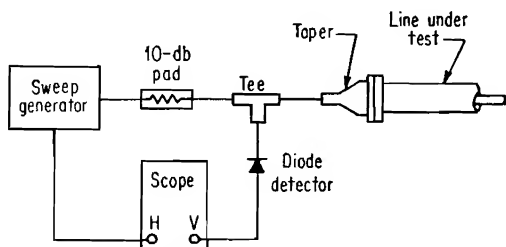


FIG. 13-69. Method of finding location of fault on line.

The impedance at the input to a transmission line will vary in a cyclic manner when there is only *one* fault on the line (173). The impedance variation can be conveniently observed by an observation of the voltage variation at the line input when the frequency is varied. The generator must, of course, be fed through a resistor, and the voltage on the generator side of the resistor must be held constant as is done in the Thévenin generators discussed earlier.

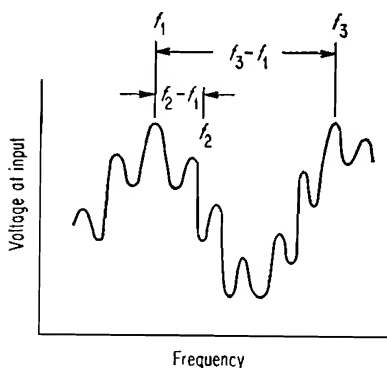


FIG. 13-70. Input voltage variation for one fault at end of line and another fault close to generator.

except that now D is the distance in feet from the generator to the fault.

It will be unusual if the antenna is a perfect match and a poor flange or bullet connection will not often give a voltage-standing-wave ratio that is any worse than the antenna. The effects of two discontinuities on the line when the input voltage is observed are shown in Figs. 13-70, 13-71, and 13-72.

Figure 13-70 shows the effect when the fault is much closer to the generator than to the antenna. Figure 13-71 shows the effect when the

generator side of the resistor must be held constant as is done in the Thévenin generators discussed earlier. If a sweep generator is used, the voltage variation can be conveniently displayed on an oscilloscope, and this will be much faster than the vswr method (174). It is not necessary to use a directional coupler, and the detector may be connected to the line input simply with a T connector as shown in Fig. 13-69. The scope display will be the same as Fig. 13-68 and Eqs. (13-33) and (13-34) will again apply

fault is about halfway along the line to the antenna, and Fig. 13-72 shows the effect when the fault is close to the antenna. In each case, $f_3 - f_1$ will give the distance *between* the faults when substituted in Eq. (13-33) or (13-34). In Figs. 13-70 and 13-71 the distance from the generator to the second fault is given by $f_2 - f_1$. In Fig. 13-72 $f_4 - f_2$ gives the distance to the fault which is causing the *larger* voltage-standing-wave ratio. If the antenna impedance varies widely over the frequency range being swept, it may be desirable to substitute a 50-ohm resistor for the antenna. If the display is still too difficult to analyze, it may be necessary to break the line in the middle and insert the 50-ohm load.

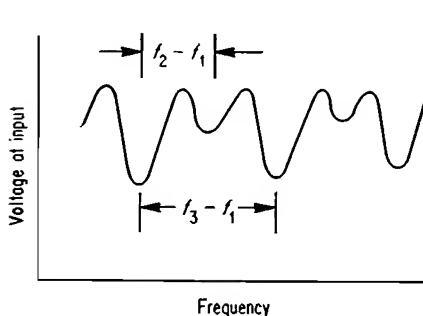


FIG. 13-71. Input voltage variation for one fault at end of line and another fault about halfway along line.

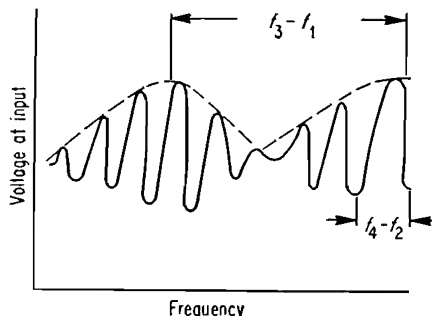


FIG. 13-72. Input voltage variation for one fault at end of line and another fault close to end of line.

Impedometer. It is possible to determine the resistive and reactive components of the load impedance by using a supplementary voltage probe in conjunction with a directional coupler. This instrument has been called an *impedometer* and is more convenient although less accurate than a slotted line (175,176).

Reflection Coefficient. The instruments connected to a directional coupler can be arranged to measure proportional instead of absolute values of power (177). The front-to-back voltage ratio is called the *reflection coefficient*, and this is generally represented by Γ . The voltage-standing-wave ratio is obtained from the reflection coefficient by

$$\text{VSWR} = \frac{1 + \Gamma}{1 - \Gamma} \quad (13-40)$$

Directional-antenna Phase Monitors. The radiation pattern of a broadcast directional antenna must be closely maintained. When the adjustments are very critical, it may be necessary to install field-strength-measuring equipment several miles out from the station with remote indicators at the station (178). In most cases it is sufficient to measure the currents to each tower and the phase angles of the radiated signals.

It is important that the phase shift in the lines to the pickup loops be known, and it is preferable that this shift be the same in each pickup line.

Methods of Measuring Phase Difference. One of the simplest but least accurate methods of measuring phase angles is to feed the two signals to the vertical and horizontal plates of an oscilloscope and to determine the phase angle from the resulting *Lissajous figure*. One serious problem with this method is that there are two angles that will produce any given figure except for the 180° figure. This ambiguity can be resolved by inserting a small phase shift in one line and observing which way the figure moves. Figure 13-73 shows how the figures change with phase angle.

There are three ways of computing the phase angle from a Lissajous figure. These are shown in Fig. 13-74 (179, 180). The scope must first be set up so that when the same signal is on both the vertical and horizontal terminals, the 0° figure is obtained. On most scopes it will be necessary to introduce a phase-shifting network in one of the input lines to make this so. The method

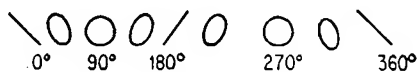


FIG. 13-73. Lissajous figures.

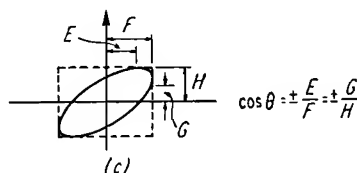
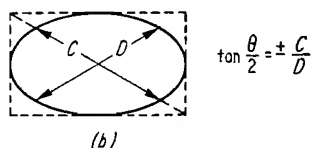
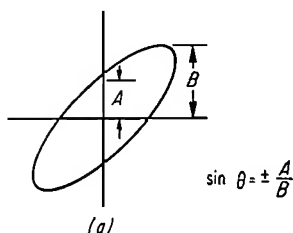


FIG. 13-74. Methods of determining angles from Lissajous figures. (Courtesy of Electronics.)

seen in Fig. 13-74a is best for angles near 0 and 180°, while the methods in parts (b) and (c) are better for angles near 90°.

A more accurate method is to insert a calibrated phase shifter in series with one of the input lines and to adjust this to obtain the 0° Lissajous figure (181). A balanced phase detector and a sensitive meter will give a more precise indication of phase and magnitude equality. A calibrated delay line can be used as a phase shifter (182). The phase shift along the delay line is obtained by simply multiplying the signal frequency by the time and 360. The calibration on the delay line can be easily checked by applying a test signal whose period is less than the full range of the delay line. The same signal is connected directly to the phase detector and also through the delay line to the phase detector. Two nulls should

be obtained as the delay line is increased, and these should be separated by the period $1/f$ of the signal. For example, if the signal is 1 Mc, the nulls should be 1 μ sec apart.

For broadcast station use, self-contained phase monitors are generally used. A number of general-purpose phase meters (183) are available.

Field-strength Measurements. Field-strength measurements are made with specialized equipment which must be accurately calibrated. The National Bureau of Standards offers a radio field-strength meter-calibration service (184). Information on field-strength meter design and operation may be found in the literature (185-192).

13-13. Spurious Transmitter Output Measurements. For best economy of the r-f spectrum a transmitter should radiate energy only on its assigned frequency and within the allocated bandwidth. All radiation which lies outside the allocated bandwidth is termed *spurious* output. The allowable spurious output from transmitters must continue to be reduced as more and more use is made of the frequency spectrum. The following types of spurious output may be found.

Harmonics. A transmitter may radiate on multiples of the carrier frequency. For the standard a-m broadcast and the 88- to 108-Mc f-m band, the FCC rules require that any emission more than 75 kc away from an a-m carrier or 600 kc away from an f-m carrier shall be attenuated at least $43 + 10 \log_{10}$ (power, in watts) db below the level of the unmodulated carrier, or 80 db, whichever is the lesser attenuation. In the case of television, the rules require an attenuation of 60 db below the visual transmitted power for any frequency more than 3 Mc above or below the channel edges.

Subharmonics. Where frequency multipliers are used in a transmitter, there is a possibility of radiation of submultiples of the carrier frequency.

In-band Noise. Spurious outputs within the frequency band of interest such as the 550- to 1,600-kc or 88- to 108-Mc broadcast bands are referred to as *in-band* noise. There are a number of sources of this type of spurious emission.

1. Overmodulation of a dsb a-m transmitter.
2. Operation of an ssb amplifier beyond linear region.
3. Keying a telegraph transmitter or pulsing a pulse transmitter with waveforms having edges that are too sharp.
4. Insufficiently suppressed mixer outputs from a frequency synthesizer.
5. Mixing action occurring at the grid of a final amplifier because of insufficient suppression of lower frequencies in frequency multiplier stages. For example, the output frequency of a 100-Mc f-m transmitter may be multiplied up from a 4-Mc oscillator. If the 4-Mc signal reaches the final grid, outputs will be generated at 96 and 104 Mc.
6. Parasitic oscillations in r-f or modulator stages.

7. R-f intermodulation because of close proximity of another transmitter. This is a particular problem in a communication center where there are a number of transmitters operating (193). For example, if transmitters operating on frequencies A and B are connected to antennas in the same general area, energy coupled back into the final amplifiers can result in generation of spurious frequencies equal to $2A - B$ or $2B - A$.

For a-m broadcast transmitters, the FCC requires 25-db attenuation from 15 to 30 kc away from the carrier and 35-db from 30 to 75 kc away from the carrier. For f-m transmitters the 25-db rule applies from 120 to 240 kc and the 35-db figure applies from 240 to 600 kc.

Cabinet Radiation. Filters in series with output transmission lines are effective in preventing harmonic power from reaching the antenna, but it is also important to keep spurious outputs from being radiated directly by the transmitter. Amateur operators have found this to be particularly important in preventing interference to nearby television receivers known as TVI (194). Complete shielding in addition to application of "electronic weatherstripping" to all cracks is frequently necessary. A gap between panels of only a few thousandths of an inch can represent a slot antenna.

Harmonic Measurement. Harmonic measurements should be made with the transmitter connected to a dummy load. It is good practice to measure all harmonics up to the tenth. It can not be assumed that the dummy load will be "flat" at the higher harmonics, and it is therefore good practice to connect a slotted line between the transmitter and dummy load. This slotted line should be at least one-half wavelength long at the second harmonic of the operating frequency. For precise measurements the coupling of the probe to the slotted line should be measured at each harmonic frequency. If a high-frequency generator is not available, it will usually be safe enough to measure the probe coupling at the operating frequency and then to assume that this coupling will increase 6 db each time the frequency is doubled. This will only be true up to the frequency at which the transmission mode in the line changes. This frequency is given by

$$f = \frac{7,520}{a + b} \quad (13-41)$$

where f = frequency, Mc

a = outside diameter of inner conductor, in.

b = inside diameter of outer conductor, in.

In fact, all harmonic measurements become very debatable above this frequency because the line may propagate in another mode at higher frequencies. In this range it will be more profitable to make actual

field-strength measurements some distance from the station, using the actual transmitting antenna.

The instrument which is connected to the slotted-line probe for making harmonic measurements may be a field-strength meter, or the arrangement shown in Fig. 13-75 may be used. Here the carrier is attenuated by band-elimination filters. These filters will pass the harmonics to the mixer where they are heterodyned with an oscillator to produce a 30-Mc

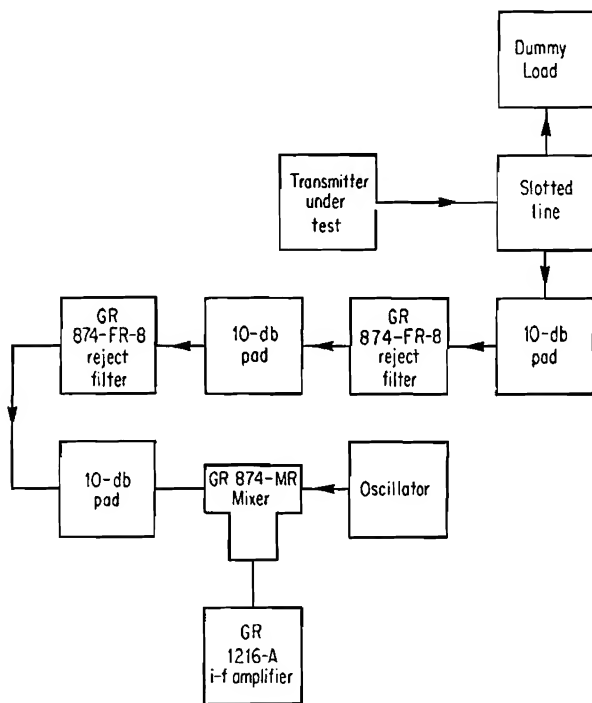


FIG. 13-75. Method of harmonic measurement. (Courtesy of General Radio Co.)

signal which is measured with an i-f amplifier. When taking a harmonic reading, the slotted-line probe is adjusted for a maximum reading in case there is a standing wave on the line at the harmonic frequency. The carrier level must be established by removing the band-elimination filters and either substituting a signal generator for the transmitter or inserting additional attenuation between the mixer and the slotted line to avoid overloading the mixer. Subharmonics will be measured in the same way by using a lower-frequency heterodyning oscillator.

In-band Noise Measurements. Spurious output measurements in the same frequency band as the transmitter can be made by means of a field-strength meter and additional band-pass or band-elimination filters to prevent the carrier from overloading the instrument. The band-

elimination filter would be tuned to the carrier, and the band-pass filter would be tuned to the frequency being measured. The filter characteristics and the selectivity of the field strength meter would determine how close to carrier the measurements can be made.

Spurious outputs very close to the carrier can be checked by connecting a sample of the transmitter output through a wideband detector to a wave analyzer. In this case the wave-analyzer readings would be referred to 0.707 times the d-c voltage reading at the detector output.

This method does not separate the upper and lower sidebands. If it is desired to obtain the upper- and lower-sideband amplitudes separately, it will be necessary to use a spectrum analyzer. Another method is to use a crystal-controlled oscillator that is operating within 200 or 250 kc of the carrier. The output of the transmitter being measured and the oscillator are then connected to a mixer which in turn is connected to a wave analyzer.

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HAZARDS ASSOCIATED WITH TRANSMITTERS

14-1. Health Hazards. Radio transmitters, particularly high-powered units, may include a number of hazards to operating personnel. Electric shock was at one time considered to be the only real hazard associated with transmitters but with most modern transmitters it is unlikely that an operator will *accidentally* come into contact with the high voltage.

With the generation of higher power levels, other hazards have become significant. Intense microwave radiation can cause damage to body tissues without any immediate sensation of heat or pain. When very high voltages are used, X rays may be radiated from rectifiers or other vacuum tubes.

Primary power lines with high current capacities are sources of burns if short-circuited. Lightning strokes can also generate high currents and high voltages. Certain insulating materials and cleaning fluids are poisonous under certain conditions.

14-2. Electric Shock. The passage of an electric current through the body has a paralyzing effect on the muscles. The effect is proportional to the current flow which is related to the voltage and resistance of contact in accordance with Ohm's law. The resistance of contact depends on many factors such as electrode configuration and size, distance between electrodes, dryness of the skin, and others. For example, there may be no sensation at all if 120 volts, 60 cps is contacted with dry calloused hands while a fatal shock may result if the skin is wet and contact area is large. The resistance between contact areas on the body can be less than 100 ohms, or more than 1 megohm, and it is impossible to give accurate figures for any particular condition. In addition, the resistances will depend on the amount of voltage applied and will become much less if the voltage is enough to puncture the skin.

The contact *area* is of particular interest. If a source of low, constant *current* is contacted, the sensation will be *less* if the contact area is *greater*. For example, the capacitors from the power line to the frame in a piece of test equipment will permit a maximum of 5 ma to flow. A grounded operator will feel this through a point contact but not if he has a good grip on the instrument. For the same reason, a person picking up a

static charge by shuffling across an insulated floor on a dry day, will feel no sensation if he discharges himself to a metallic ground through a coin. In other words, the contact area of two fingers on a dime will reduce the *milliamperes per square centimeter* to a value less than that producing a sensation.

On the other hand, a constant-voltage source such as a battery, regulated power supply, or main power line will cause a *greater* sensation if the contact area is *greater*. There will also be a much greater effect from a constant-voltage source if the voltage is sufficient to break down the skin, regardless of the contact area.

Current flows of 1 to 5 ma produce little or no sensation depending on the area of contact. It has been reported that currents of 15 to 20 ma through muscles cause paralysis and that currents of 75 to 100 ma through the heart may be fatal (1,2). These values actually have little practical meaning because the skin contact resistance varies over such wide limits that the current resulting from contact with a given voltage can not be predicted.

The effects depend on frequency, and the peak danger occurs at about 60 cps. Direct contact with r-f voltages will cause surface burns but the current will drop rapidly with penetration.

The maximum allowable value of leakage current from the frame of an electrical appliance or test instrument to ground is 5 ma. This leakage current is approached when 0.1- μ f capacitors are connected between the power line and frame in a 117-volt 60-cps instrument. Many people find this much current unpleasant and it is good practice to fit all radio-station outlets with grounding-type receptacles so that all test instruments may have the frames grounded through the three-pin grounding plugs.

14-3. Treatment of Electric Shock. It is important to remember that immediate action must be taken in electric shock cases. The brain will seldom recover "completely" after more than $3\frac{1}{2}$ to 4 min of cessation of active blood flow. In other words, the victim may be revived but remain blind or deaf or suffer other permanent malfunction. It cannot be assumed, however, that this cessation has occurred, and the immediate task is to ensure a supply of oxygen to the lungs. It is for these reasons that artificial respiration is administered to power-company lineman at the "top of the pole" when they suffer a disabling shock.

Professional help should also be summoned without delay. The fire department will almost always be the fastest available assistance and they usually have resuscitation equipment. Most police departments are organized to respond to fire calls, and they in turn can summon medical help most effectively.

When a victim suffering from electric shock is found, the important steps to take are:

1. *Pull the main switch.*
2. *Blow 12 times per minute into his mouth.*
3. *Have someone call the fire department.*

Artificial Respiration. There are at least four methods of artificial respiration, some of which can be quite dangerous in the hands of over-enthusiastic rescuers. It is important to avoid wasting time arguing which method to use because the treatment *must* be started within 3 or 4 min of the time that breathing has stopped. None of the methods are as good as the use of resuscitation equipment and this is why the fire department should be immediately called.

One practical demonstration from a first aid instructor is more effective than diagrams or printed instructions, and everyone working with high voltage apparatus should be instructed in the mouth-to-mouth technique.

There is a range of current which may cause heart stoppage. A technique called *closed-chest cardiac massage* is sometimes successful in maintaining blood flow (3). With the victim lying face up on a rigid support, the operator places the heel of his hand about 2 inches above the lower end of the breastbone (sternum). The other hand is placed over the first and the operator alternately exerts pressure on the chest with his full body weight and then releases pressure about 60 times per minute. Breathing assistance at 12 times per minute should also be continued.

14-4. Precautions against Electric Shock. The principal precautions that can be taken against accidental contact with high voltages are:

1. Power-interlock switches on cabinet doors
2. Grounding switches controlled by cabinet doors
3. Grounding sticks
4. Bleeder resistors across capacitors and nonlinear resistors (thyrites or varistors) from one side of capacitor to ground

Door Interlocks. The simplest door-interlock switches are similar to the type used on refrigerator or automobile doors for operating lights. In some cases it is desirable to make the interlock switches "cheatable." These switches are momentary-contact when pushed and snap-action when pulled. Each time the door is closed, the switch is restored to the "noncheated" condition. In certain types of military equipment a switch, called a *battle short*, is provided to short out all door interlocks.

Switch Locks. A device sometimes used to ensure that power is switched off before transmitter doors are opened is a lock on the main power switch. The key or keys required to unlock the transmitter doors can only be released when the main power switch is locked open.

Grounding Switches. As a precaution against open capacitor-bleeder resistors, gravity-operated grounding switches are frequently used to discharge capacitors operating at more than 1,000 volts. These are generally mechanically operated by the doors; but in a large installation, one grounding switch may be electrically controlled from all doors. It is important that heavy wiring be run directly from the switch to the capacitors.

Grounding Sticks. As a final precaution before touching any parts inside a transmitter, an operator should ground high-voltage components with a grounding stick. Insulated wire for the ground connection is not recommended for the ground connection because of the danger of wire breakage inside the insulation. The recommended connection is with sash chain. It is desirable to provide an interlocked hook switch to hang the stick on, so that the transmitter cannot be switched on with the high voltage grounded.

Bleeder Resistors. High-voltage capacitors can store sufficient energy to cause serious shocks even when the capacity is less than $0.1 \mu\text{f}$. Bleeder resistors should be connected across all high-voltage capacitors, and if these are wire-wound it is good practice to either use a parallel string or to back up the wire-wound units with higher-value carbon resistor units.

In some transmitter circuits metering circuits make it inconvenient to directly ground either the positive or negative terminals of the high-voltage supply. In these cases, the side which operates nearest ground

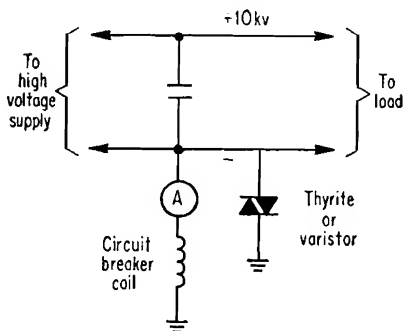


FIG. 14-1. Protection of negative side of high-voltage supply with nonlinear resistance.

should have a thyrite or varistor connected between it and ground. These nonlinear resistors will carry negligible current at the operating voltage, but their resistance will drop rapidly if the voltage is increased. This will prevent the end of the capacitor, which is normally at a low potential, from moving to a high potential in the event a solid short to ground should occur at the high-potential side of the supply. Figure 14-1 illustrates a typical circuit.

Meters and circuit-breaker coils are usually operated as close to ground potential as possible but may introduce appreciable resistance and inductance. In the circuit shown in Fig. 14-1, if the meter and breaker coil have a resistance of 1 ohm and if a ground fault of 0.1 ohm occurs from the positive terminal to ground, the negative terminal will rise to $10/11$ of the output voltage unless it is "held down" with a nonlinear resistor.

14-5. Lightning Hazards. Another possible source of electric shock is the voltage that may be set up when a station is struck by lightning. A 250-ft radio tower located in a normal area which has 30 to 35 thunderstorm days per year will average one direct hit per year (4). The Empire State Building in New York City averages 23 direct hits per year (5). Stroke currents will vary from 5,000 to 50,000 amp, although currents of over 300,000 amp have been recorded (6). The voltage developed at any point of interest will be the current multiplied by the *surge* impedance between that point and ground. This means that ground connections should not only have minimum resistance but also must have minimum inductance. Sheet copper, several inches wide, will provide good low-resistance, low-inductance connections. Transmission-line outer conductors, waveguide, and support structures should be adequately grounded before entering a transmitter building. A suitable ground can be made by driving pipes 10 ft long into the ground. These should circle the building and be placed 6 to 10 ft apart. The pipes should be interconnected with wire which is AWG no. 6 or larger, and this wire should be buried about 1 ft into the earth. The grounding can sometimes be improved by pouring rocksalt into the pipes and then pouring in water to distribute it at the bottom.

At installations where the smaller transmission lines are used, it is very important that a parallel path be provided for high currents. Coaxial lines can be crushed by magnetic forces directed radially inwards when high currents flow in the outer conductor.

Inside the station it is important to make sure that equipment handled by operators cannot "move away" from each other in potential during a direct hit. For example, an operator holding a telephone and touching a control console could be shocked if the telephone ground and console ground move apart in potential during a hit.

There is a considerable amount of literature on the subject of lightning protection (7-11). In the case of a standard broadcast a-m tower it is usually not possible to ground it through a low impedance because of the necessity of tuning it to the operating frequency. In this case spark gaps must be installed at the base but precautions must be taken to avoid a "following power arc." Once the air is ionized by the lightning flash, the transmitter power may maintain the arc. This can be avoided by installing a relay to cut off the high voltage to the final transmitter stage whenever an arc occurs (12).

14-6. Power-line Shorts. The dangers associated with short-circuiting a high-capacity power line should be recognized. Operators wearing rings or metal watchbands should keep their hands away from an energized high-current line even if it is only operating at 5 volts.

Small-size conductors should never be bridged across a heavy conductor

line without first going through fuses or a circuit breaker. The circuit breaker should have an interrupting capacity greater than what the line can supply *before* the next breaker along the line trips. Power-line short circuits are limited only by reactance in the line and distribution transformers during the period between the occurrence of a short and opening of the fuse or breaker. If the fuse is operated beyond its *voltage* rating or the breaker beyond its *interrupting capacity*, an arc may be maintained across the fuse or breaker contacts after they have "opened."

14-7. X-ray and R-F Radiation Hazards. Exposure of the body to electromagnetic radiation does not become a serious problem until *irreversible* effects take place. Such effects can be induced by both r-f and X radiation. Exposure to intense r-f radiation produces a purely thermal effect; but this can cause the eye to develop opacities of the lenses (cataracts), and these can be irreversible. X radiation can damage the eye lenses and can also cause irreversible damage to any cell in the body (13). Sometimes the cells will recover, but the damaged cells can become self-regenerating. Damage to cells of the reproductive tissues is called *genetic*, and damage to other cells is called *somatic*. The somatic damage has been known to become malignant (cancer or leukemia) or, in milder form, it may appear as an apparent acceleration of the aging process.

R-f Radiation. The amount of r-f radiation that can be tolerated is expressed as a power *density* and is usually given in milliwatts per square centimeter. The highest densities for a given power level will occur in narrow microwave beams. It must not be assumed, however, that there is no danger at the longer wavelengths, and work should not be performed on an energized a-m broadcast antenna tower.

Radio-frequency energy may penetrate as much as one-tenth wavelength at microwaves, but at longer wavelengths the penetration may be much less than 1/1,000 of a wavelength (14).

The amount of power absorbed by the body varies with frequency, and consequently the tolerable power density varies (15,16).

In one set of experiments, an eye cataract was developed in an animal after 90 min of exposure at a power density of 290 mw per sq cm. The safe level of exposure is obviously much less than this.

It should be noted that a person wearing eyeglasses is not protected by looking away from the radiation because eyeglasses can actually focus r-f radiation into the eye by reflection from the wearer's side of the glass.

A maximum safe level of 10 mw per sq cm has been suggested (21). For a situation such as over-the-horizon stations where operators are constantly exposed considerably lower levels have been recommended. Commercial radiation detectors are available which give readings directly in terms of milliwatts per square centimeter. It has been reported that

NE-51 neon bulbs without anything connected to them will fire when the intensity of microwave radiation is 5 to 6 mw per sq cm, while photographic flash bulbs will discharge at a level of 1 to 2 mw per sq cm (17).

The power density or *flux* at a point in the "distant field" in front of an antenna can be estimated from the relation (18)

$$\delta_{avg} = \frac{2,000P}{D^2} \left(\frac{R}{r} \right)^2 \quad (14-1)$$

where δ_{avg} = average power density, mw per sq cm

P = power delivered to antenna by transmission line, kw

D = antenna diameter, ft

$R = D^2/2\lambda$ (Rayleigh or Fresnel region)

λ = wavelength, ft

r = distance in front of antenna, ft

Equation (14-1) is true for all distances greater than about twice the Rayleigh distance R from the antenna. The absolute maximum power flux occurs at about $0.5R$ from the antenna and is about $3860P/D^2$.

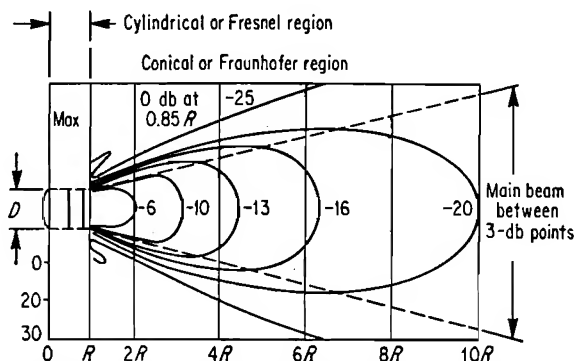


FIG. 14-2. Radiation field in front of parabolic reflector. (Courtesy of Nature.)

To obtain the average power in front of an antenna which is connected to a radar or other type of pulse transmitter, Eq. (14-1) should be multiplied by the duty ratio.

The power-density distribution in front of a parabolic reflector is illustrated in Fig. 14-2. The contours are given in decibels referred to the power density at the Rayleigh distance. The values given by Eq. (14-1) and the contours shown in Fig. 14-2 are for a theoretical condition in free space; and since there will be ground reflections in a practical situation, a safety factor of 4 should be allowed.

There may be danger areas in the immediate vicinity of the transmitter. If an operator suspects an arc or a flash inside a transmission line, waveguide, or power-amplifier cavity he should never attempt to

make a small opening to look inside with the power on. All joints should be closed as tightly as possible.

A section of "waveguide beyond cutoff" is sometimes used for air-cooling purposes in the output cavities of high-powered amplifiers; and although r-f radiation should be negligible through these tubes, there is nothing to stop X rays, and an operator should not hold his eye close to one of these openings.

Even with relatively low-powered transmitters, dangerous power densities are present over the cross section of the waveguide. For example, microwave transmitters used in the 4-kMc band have an output of only 0.5 watt. It is usual to connect six of these transmitters to a single waveguide run which has a cross section of 17 sq cm. The power density could be as much as 177 mw per sq cm, which is close to 20 times greater than the safe level. The highest temperatures produced in the 4-kMc frequency range can be expected to occur about 1 cm below the skin surface, and the hazard to eyesight is obvious. The material within the eye lens is similar to the white of an egg which is one of the substances known to become more viscous and to harden with a temperature increase.

An operator should never look into an energized waveguide. It should be remembered that there is a small area behind most reflectors where energy coming directly out of the feed horn can graze past the edges of the reflector. In a typical 1,000-Mc installation the gain of the feed horn will just about make up for the difference between dish-edge and dish-center illumination. The power density at a distance from the horn will then be given by

$$\delta = \frac{84.5P}{r^2} \quad (14-2)$$

where δ = power density, mw per sq cm

P = horn output power, kw

r = distance from horn, ft

The danger area around transmitter output circuits can be measured with a radiation detector or can be estimated from a field-strength measurement as given by the relation

$$\delta = \frac{E^2}{3,770} \quad (14-3)$$

where δ = power density, mw per sq cm

E = volts per m

The danger area, assuming a 10 mw per sq cm flux, then starts at about 194 volts per m.

X Rays. X rays may be generated in a radio transmitter wherever high-energy electrons strike a target. The energy that an electron

acquires after passing through a potential difference of 1 volt is called an *electron volt*. Particle energies are usually seen expressed in thousands (kev), millions (mev), or billions (bev) of electron volts. In high-voltage electron tubes such as hydrogen thyratrons and klystrons, X rays may be produced when electrons emitted from the grid or cathode are accelerated towards the anode. Observations on high-voltage thyratrons showed that X rays emanated in a narrow beam in the horizontal direction from the grid-anode region (19). X rays were observed when the anode voltage was 18 kv and above. The amount of X radiation will depend upon both the voltage and the current.

Genetic and somatic effects are accumulative, but most other effects of X radiation are not accumulative. The accumulative effects place an upper limit on the total radiation dose, and this is measured in *roentgens*. Portable radiation indicators are calibrated in terms of the *dosage rate* which is usually read in *milliroentgens per hour* (mr per hr). The total dose is then determined by multiplying the rate by the number of hours of exposure.

The radiation may not be all absorbed in the tissue at the site of interest, and another unit called the *rad* (radiation absorbed dose) refers to the amount of radiation actually absorbed (20). Rads and roentgens are roughly equivalent in quantity. A *rep* (roentgen equivalent physical) refers to radiation other than X rays which will cause an absorption equal to 1 roentgen. Different sources of radiation have different effects which are related to X rays by a number which is referred to as the *rbe* (relative biological effectiveness). The rbe dose is expressed in *rems* (roentgen equivalent man) or *rebs* (roentgen equivalent biological).

$$\text{rems} = \text{rbe} \times \text{rads} \quad (14-4)$$

Alpha particles and fast neutrons have an rbe of 10, but any radiation produced by a radio transmitter will have an rbe of 1.

Permissible X-ray Exposures. Radiation exposures which are considered to be safe have constantly dropped since the discovery of X rays. In 1957 the National Committee on Radiation Protection and Measurements specified a maximum accumulated permissible dose (MPD) to the whole body as

$$\text{MPD} = 5(N - 18) \quad (14-5)$$

where MPD = rems

N = person's age

The maximum total dose in any one 13-week period was limited to 3 rems. In 1959 an international committee recommended that the maximum dose for atomic-energy workers be limited to 5 rems per year. They also recommended that the limit for areas in the vicinity of an atomic energy installation be set at one-tenth of this amount or 0.5 rem per year.

There is enough radium in an average luminous wristwatch to radiate 4 mr.

The U.S. Air Force has specified that the maximum radiation from ground electronic equipment must not exceed 100 mr per week (21).

The observations discussed in Ref. 19 show a measurement of 1,230 mr per hr at a distance of 1 ft from a type 1257 hydrogen thyratron operating at 42 kv. X rays developed at this voltage level are relatively "soft" and can be reduced to a negligible quantity with $\frac{1}{4}$ in. of leaded glass or $\frac{1}{16}$ in. of steel paneling (22,23).

There are a number of photographic and electronic methods of measuring radiation intensity (24-26).

14-8. Chemical Hazards. Most chemical hazards associated with radio transmitters occur as a result of failure or overheating of a component.

Insulating Oils. It is necessary to use nonflammable oils in transformers, capacitors, and other components which do not operate in a concrete vault. These chemicals are usually chlorinated biphenyls. Toxic fumes are liberated when this material is heated, and the maximum safe concentration is 1.0 to 2.0 mg per cu m of air. Transformers and chokes fitted with "breathers" should be vented to the outside. The liquid can be absorbed through the skin; so hands should not be immersed unless they can be washed fairly quickly afterward.

Teflon. Teflon insulation releases vapors when heated to a point above 400°F. It is not particularly dangerous, but inhalation can cause an attack similar to influenza which has been called *polymer fume fever*. Symptoms do not occur until several hours after exposure, and the attack passes after 36 to 48 hr.

Selenium. A selenium rectifier which has been overheated should be approached with caution. The allowable concentration of selenium is 0.1 ppm in air. The material resembles arsenic both chemically and physiologically. It causes respiratory irritation, abdominal pain, and irritation to the brain (27).

Carbon Tetrachloride. Carbon tetrachloride is frequently used for maintenance purposes and is a dangerous chemical at room temperature. The maximum allowable concentration in air is 25 ppm. It causes eye inflammation and damage to the liver, kidneys, heart, and nervous system. Absorption of 3 to 4 ml can be fatal. The toxic symptoms may not appear until 2 to 8 days after exposure.

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APPENDIX

TABLE A-1. INTERNATIONAL CALL SIGN PREFIXES*

AAA-ALZ	United States of America	JZA-JZZ	Netherlands New Guinea
AMA-AOZ	Spain	KAA-KZZ	United States of America
APA-ASZ	Pakistan	LAA-LNZ	Norway
ATA-AWZ	India	LOA-LWZ	Argentine Republic
AXA-AXZ	Commonwealth of Australia	LXA-LXZ	Luxembourg
AYA-AZZ	Argentine Republic	LYA-LYZ	Lithuania
BAA-BZZ	China	LZA-LZZ	People's Republic of Bulgaria
CAA-CEZ	Chile	MAA-MZZ	Great Britain
CFA-CKZ	Canada	NAA-NZZ	United States of America
CLA-CMZ	Cuba	OAA-OCZ	Peru
CNA-CNZ	Morocco	ODA-ODZ	Lebanon
COA-COZ	Cuba	OEA-OEZ	Austria
CPA-CPZ	Bolivia	OFA-OJZ	Finland
CQA-CRZ	Portuguese Overseas Provinces	OKA-OMZ	Czechoslovakia
CSA-CUZ	Portugal	ONA-OTZ	Belgium and Colonies
CVA-CXZ	Uruguay	OUA-OZZ	Denmark
CYA-CZZ	Canada	PAA-PIZ	Netherlands
DAA-DMZ	Germany	PJA-PJZ	Netherlands Antilles
DNA-DQZ	Belgian Congo	PKA-POZ	Republic of Indonesia
DRA-DTZ	Bielorussian Soviet Socialist Re- public	PPA-PYZ	Brazil
DUA-DZZ	Republic of the Philippines	PZA-PZZ	Surinam
EAA-EHZ	Spain	QAA-QZZ	(Service abbreviations)
EIA-EJZ	Ireland	RAA-RZZ	Union of Soviet Socialist Republics
EKA-EKZ	Union of Soviet Socialist Republics	SAA-SMZ	Sweden
ELA-ELZ	Liberia	SNA-SRZ	People's Republic of Poland
EMA-EOZ	Union of Soviet Socialist Republics	SSA-SSM	Egypt
EPA-EQZ	Iran	SSN-STZ	Sudan
ERA-ERZ	Union of Soviet Socialist Republics	SUA-SUZ	Egypt
ESA-ESZ	Estonia	SVA-SZZ	Greece
ETA-ETZ	Ethiopia	TAA-TCZ	Turkey
EUA-EZZ	Union of Soviet Socialist Republics	TDA-TDZ	Guatemala
FAA-FZZ	France and Colonies and Protector- ates	TEA-TEZ	Costa Rica
GAA-GZZ	Great Britain	TFA-TFZ	Iceland
HAA-HAZ	Hungarian People's Republic	TGA-TGZ	Guatemala
HBA-HBZ	Switzerland	THA-THZ	France and Colonies and Protector- ates
HCA-HDZ	Ecuador	TIA-TIZ	Costa Rica
HEA-HEZ	Switzerland	TJA-TZZ	France and Colonies and Protector- ates
HFA-HFZ	People's Republic of Poland	UAA-UQZ	Union of Soviet Socialist Republics
HGA-HGZ	Hungarian People's Republic	URA-UTZ	Ukrainian Soviet Socialist Republic
HHA-HHZ	Republic of Haiti	UUA-UZZ	Union of Soviet Socialist Republics
HIA-HIZ	Dominican Republic	VAA-VGZ	Canada
HJA-HKZ	Republic of Colombia	VHA-VNZ	Commonwealth of Australia
HLA-HMZ	Korea	VOA-VOZ	Canada
HNA-HNZ	Iraq	VPA-VSZ	British Colonies and Protectorates
HOA-HPZ	Republic of Panama	VTA-VWZ	India
HQA-HRZ	Republic of Honduras	VXA-VYZ	Canada
HSA-HSZ	Thailand	VZA-VZZ	Commonwealth of Australia
HTA-HTZ	Nicaragua	WAA-WZZ	United States of America
HUA-HUZ	Republic of El Salvador	XAA-XIZ	Mexico
HVA-HVZ	Vatican City State	XJA-XOZ	Canada
HWA-HYZ	France and Colonies and Protector- ates	XPA-XPZ	Denmark
HZA-HZZ	Saudi Arabia	XQA-XRZ	Chile
IAA-IZZ	Italy and Colonies	XSA-XSZ	China
JAA-JSZ	Japan	XTA-XTZ	France and Colonies and Protector- ates
JTA-JVZ	Mongolian People's Republic	XUA-XUZ	Cambodia
JWA-JXZ	Norway	XVA-SVZ	Viet-Nam
JYA-JYZ	Jordan	XWA-XWZ	Laos

* Courtesy of American Radio Relay League.

TABLE A-1. INTERNATIONAL CALL SIGN PREFIXES* (*Continued*)

XXA-XXZ	Portuguese Overseas Provinces	3HA-3UZ	China
XVA-XZZ	Burma	3VA-3VZ	Tunisia
YAA-YAZ	Afghanistan	3WA-3WZ	Viet-Nam
YBA-YHZ	Republic of Indonesia	3YA-3YZ	Norway
YIA-YIZ	Iraq	3ZA-3ZZ	People's Republic of Poland
YJA-YJZ	New Hebrides	4AA-4CZ	Mexico
YKA-YKZ	Syrian Republic	4DA-4IZ	Republic of the Philippines
YLA-YLZ	Latvia	4JA-4LZ	Union of Soviet Socialist Republics
YMA-YMZ	Turkey	4MA-4MZ	Venezuela
YNA-YNZ	Nicaragua	4NA-4OZ	Yugoslavia
YOA-YRZ	Roumanian People's Republic	4PA-4SZ	Ceylon
YSA-YSZ	Republic of El Salvador	4TA-4TZ	Peru
YTA-YUZ	Yugoslavia	4UA-4UZ	United Nations
YVA-YYZ	Venezuela	4VA-4VZ	Republic of Haiti
YZA-YZZ	Yugoslavia	4WA-4WZ	Yemen
ZAA-ZAZ	Albania	4XA-4XZ	State of Israel
ZBA-ZJZ	British Colonies and Protectorates	4YA-4YZ	International Civil Aviation Organization
ZKA-ZMZ	New Zealand	5AA-5AZ	Libya
ZNA-ZOZ	British Colonies and Protectorates	5CA-5CZ	Morocco
ZPA-ZPZ	Paraguay	5LA-5LZ	Liberia
ZQA-ZQZ	British Colonies and Protectorates	5PA-5QZ	Denmark
ZRA-ZUZ	Union of South Africa	9AA-9AZ	San Marino
ZVA-ZZZ	Brazil	9KA-9KZ	Kuwait
2AA-2ZZ	Great Britain	9NA-9NZ	Nepal
3AA-3AZ	Monaco	9SA-9SZ	Saar
3BA-3FZ	Canada		
3GA-3GZ	Chile		

* Courtesy of American Radio Relay League.

TABLE A-2. UNITED STATES CALL SIGN ALLOCATIONS

Class of station (1)	Composition of call sign (2)	Call sign blocks available (3)
Coast.....	3 letters	KAA to KZZ. WAA to WZZ.
Aeronautical.....	3 letters, 1 digit	KAA2 to KZZ9. WAA2 to WZZ9.
Fixed coastal telephone in Alaska.....	3 letters, 2 digits	KAA20 to KZZ99. WAA20 to WZZ99.
Land (other than aeronautical and coast)...	3 letters, 3 digits	KAA200 to KZZ999. WAA200 to WZZ999.
Mobile telegraph (other than ship and aircraft).	4 letters, 1 digit	KAAA2 to KZZZ9.
Mobile telephone (other than ship and aircraft).	2 letters, 4 digits	WAAA2 to WZZZ9.
Ship telegraph.....	4 letters	KA2000 to KZ9999.
Ship telephone.....	2 letters, 4 digits	KAAA to KZZZ.
Ship telegraph and telephone.....	Same as for ship telegraph	WAAA to WZZZ.
Ship radar.....	Same as ship telephone	WA2000 to WZ9999.
Ship radar and telegraph.....	Same as for ship telegraph	
Ship radar and telephone.....	Same as for ship telephone	
Aircraft telegraph.....	5 letters	KAAAA to KYZZZ. WAAA to WZZZZ.
Aircraft telephone.....	Registration Number	
Aircraft telegraph and telephone.....	Same as for Aircraft telegraph	
Lifeboats, liferafts and other survival craft	Call sign of parent ship or Aircraft plus 2 digits from 20 to 99 inclusive. The parent call sign must in such cases be a 5 letter call, if aircraft, or a 4 letter call, if a ship.	
Broadcasting (standard).....		KAAA to KZZZ. WAAA to WZZZ.
Broadcasting (f-m).....	4 letters	KAAA to KZZZ. WAAA to WZZZ.
Broadcasting (f-m) (where the last 2 letters are f-m).	5 letters	KAA-FM to KZZ-FM. WAA-FM to WZZ-FM.
	6 letters	KAAA-FM to KZZZ-FM. WAAA-FM to WZZZ-FM.
Broadcasting (television).....	4 letters	KAAA to KZZZ. WAAA to WZZZ.
Broadcasting (television) (where the last 2 letters are TV).	5 letters	KAA-TV to KZZ-TV. WAA-TV to WZZ-TV.
	6 letters	KAAA-TV to KZZZ-TV. WAAA-TV to WZZZ-TV.
Experimental (where the letter X follows the digit).	(2 letters, 1 digit, 3 letters)	KA2XAA to KZ9XZZ. WA2XAA to WZ9XZZ.
Amateur (letter X may not follow digit)...	(1 letter, 1 digit, 2 letters)	K1AA to KØZZ. W1AA to WØZZ.
	(1 letter, 1 digit, 3 letters)	K1AAA to KØZZZ. W1AAA to WØZZZ.
	(2 letters, 1 digit, 2 letters)	KA1AA to KZØZZ. WA1AA to WZØZZ.
	(2 letters, 1 digit, 3 letters)	KA1AAA to KZØZZZ. WA1AAA to WZØZZZ.
Standard frequency.....		WW1, WWV, WWVH, and WWVL.

TABLE A-2. UNITED STATES CALL SIGN ALLOCATIONS (*Continued*)

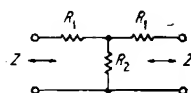
<i>Call sign area</i>	<i>Call sequence*</i>
Colorado, Iowa, Kansas, Minnesota, Missouri, Nebraska, North Dakota, South Dakota.....	KAA-KBZ WAA-WBZ
Connecticut, Maine, Massachusetts, New Hampshire, Rhode Island, Vermont.....	KCA-KDZ WCA-WDZ
New Jersey, New York.....	KEA-KFZ WEA-WFZ
Delaware, District of Columbia, Maryland, Pennsylvania.....	KGA-KHZ WGA-WHZ
Alabama, Georgia, Florida, Kentucky, North Carolina, South Carolina, Tennessee, Virginia.....	KIA-KJZ WIA-WJZ
Arkansas, Louisiana, Mississippi, New Mexico, Oklahoma, Texas.....	KKA-KLZ WKA-WLZ
California.....	KMA-KNZ WMA-WNZ
Arizona, Idaho, Montana, Nevada, Oregon, Utah, Washington, Wyoming.....	KOA-KPZ WOA-WPZ
Michigan, Ohio, West Virginia.....	KQA-KRZ WQA-WRZ
Illinois, Indiana, Wisconsin.....	KSA-KTZ WSA-WTZ
Pacific areas.....	KUA-KVZ
Alaska.....	KWA-KZZ
Atlantic-Caribbean areas.....	WWA-WWZ

* Broadcasting station calls do not follow the sequence of this table. The first two letters of Experimental Station calls will follow the sequence of this table.

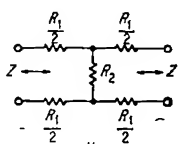
TABLE A-3. SYMMETRICAL T, H, π , OR O ATTENUATORS*

To obtain resistance values, multiply the required impedance by the numbers shown.

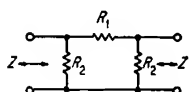
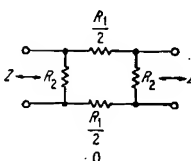
Attenuation, db	T or H		π or O	
	Series R_1 Z	Shunt R_2 Z	Series R_1 Z	Shunt R_2 Z
0.2	0.0116	43.40	0.0230	86.80
0.4	0.0230	21.60	0.0460	43.40
0.6	0.0346	14.46	0.0692	29.00
0.8	0.0460	10.82	0.0922	21.74
1	0.0575	8.66	0.114	17.39
2	0.1145	4.30	0.232	8.73
3	0.1715	2.84	0.353	5.83
4	0.2265	2.10	0.477	4.415
5	0.2800	1.65	0.605	3.571
6	0.3320	1.34	0.746	3.012
7	0.3830	1.12	0.893	2.611
8	0.4310	0.947	1.058	2.320
9	0.4760	0.812	1.228	2.100
10	0.5200	0.702	1.425	1.923
11	0.5600	0.618	1.615	1.786
12	0.5980	0.536	1.865	1.672
13	0.6350	0.472	2.120	1.575
14	0.6670	0.416	2.405	1.500
15	0.6980	0.367	2.73	1.433
16	0.7260	0.325	3.08	1.379
17	0.7530	0.288	3.47	1.328
18	0.7770	0.256	3.91	1.290
19	0.7940	0.233	4.29	1.260
20	0.8180	0.202	4.95	1.222
22	0.8540	0.160	6.25	1.171
24	0.8810	0.1265	7.90	1.136
26	0.9050	0.1005	9.95	1.105
28	0.9230	0.0798	12.52	1.083
30	0.9390	0.0632	15.85	1.065
32	0.9500	0.0502	19.95	1.052
34	0.9620	0.0400	25.00	1.040
36	0.9690	0.0317	31.6	1.032
38	0.9750	0.0252	39.7	1.026
40	0.9800	0.0200	50.0	1.020
42	0.9850	0.0159	63.0	1.015
44	0.9880	0.01262	79.2	1.012
46	0.9900	0.01002	99.7	1.010
48	0.9920	0.00795	125.9	1.008
50	0.9940	0.00631	158.5	1.006



T



H

 π 

O

* From "Reference Data for Radio Engineers," International Telephone & Telegraph Corporation.

TABLE A-4. DECIBELS AND POWER, VOLTAGE, AND CURRENT RATIOS*

Power ratio	Voltage and current ratio	Decibels†	Power ratio‡	Voltage and current ratio	Decibels
1.0233	1.0116	0.1	19.953	4.4668	13.0
1.0471	1.0233	0.2	25.119	5.0119	14.0
1.0715	1.0351	0.3	31.623	5.6234	15.0
1.0965	1.0471	0.4	39.811	6.3096	16.0
1.1220	1.0593	0.5	50.119	7.0795	17.0
1.1482	1.0715	0.6	63.096	7.9433	18.0
1.1749	1.0839	0.7	79.433	8.9125	19.0
1.2023	1.0965	0.8	100.00	10.0000	20.0
1.2303	1.1092	0.9	158.49	12.589	22.0
1.2589	1.1220	1.0	251.19	15.849	24.0
1.3183	1.1482	1.2	398.11	19.953	26.0
1.3804	1.1749	1.4	630.96	25.119	28.0
1.4454	1.2023	1.6	1000.0	31.623	30.0
1.5136	1.2303	1.8	1584.9	39.811	32.0
1.5849	1.2589	2.0	2511.9	50.119	34.0
1.6595	1.2882	2.2	3981.1	63.096	36.0
1.7378	1.3183	2.4	6309.6	79.433	38.0
1.8197	1.3490	2.6	10 ⁴	100.000	40.0
1.9055	1.3804	2.8	10 ⁴ × 1.5849	125.89	42.0
1.9953	1.4125	3.0	10 ⁴ × 2.5119	158.49	44.0
2.2387	1.4962	3.5	10 ⁴ × 3.9811	199.53	46.0
2.5119	1.5849	4.0	10 ⁴ × 6.3096	251.19	48.0
2.8184	1.6788	4.5	10 ⁵	316.23	50.0
3.1623	1.7783	5.0	10 ⁵ × 1.5849	398.11	52.0
3.5481	1.8836	5.5	10 ⁵ × 2.5119	501.19	54.0
3.9811	1.9953	6.0	10 ⁵ × 3.9811	630.96	56.0
5.0119	2.2387	7.0	10 ⁵ × 6.3096	794.33	58.0
6.3096	2.5119	8.0	10 ⁶	1,000.00	60.0
7.9433	2.8184	9.0	10 ⁷	3,162.3	70.0
10.0000	3.1623	10.0	10 ⁸	10,000.0	80.0
12.589	3.5481	11.0	10 ⁹	31,623	90.0
15.849	3.9811	12.0	10 ¹⁰	100,000	100.0

* From "Reference Data for Radio Engineers," International Telephone & Telegraph Corporation.

† To convert decibels to nepers multiply by 0.1151; nepers to decibels multiply by 8.686.

‡ Where the power ratio is less than unity, it is usual to invert the fraction and express the answer as a decibel loss.

TABLE A-5. DECIBEL REFERENCE LEVELS

0 dbm.....	1 mw
0 dbw.....	1 watt
0 dbk.....	1 kw
0 db μ	1 μ volt per m
0 dbv.....	1 volt
0 dba.....	-85 dbm (F1A weighted)*

Example: 500 watts = +57 dbm = +27 dbw = -3 dbk

* If the interfering noise is known to be evenly distributed across the 3,000-cps voice frequency band, the reference level is -82 dbm.

TABLE A-6. ANNEALED COPPER (AWG)*

AWG B & S gauge	Diam- eter, mils	Cross section		Ohms per 1,000 ft at 20°C (68°F)†	Lb per 1,000 ft	Ft per lb	Ft per ohm at 20°C (68°F)	Ohms per lb at 20°C (68°F)
		Cir mils	Sq in.					
0000	460.0	211,600	0.1662	0.04901	640.5	1.561	20,400	0.00007652
000	409.6	167,800	0.1318	0.06180	507.9	1.968	16,180	0.0001217
00	364.8	133,100	0.1045	0.07793	402.8	2.482	12,830	0.0001935
0	324.9	105,500	0.08289	0.09827	319.5	3.130	10,180	0.0003076
1	289.3	83,690	0.06573	0.1239	253.3	3.947	8,070	0.0004891
2	257.6	66,370	0.05213	0.1563	200.9	4.977	6,400	0.0007778
3	229.4	52,640	0.04134	0.1970	159.3	6.276	5,075	0.001237
4	204.3	41,740	0.03278	0.2485	126.4	7.914	4,025	0.001966
5	181.9	33,100	0.02600	0.3133	100.2	9.980	3,192	0.003127
6	162.0	26,250	0.02062	0.3951	79.46	12.58	2,531	0.004972
7	144.3	20,820	0.01635	0.4982	63.02	15.87	2,007	0.007905
8	128.5	16,510	0.01297	0.6282	49.98	20.01	1,592	0.01257
9	114.4	13,090	0.01028	0.7921	39.63	25.23	1,262	0.01999
10	101.9	10,380	0.008155	0.9989	31.43	31.82	1,001	0.03178
11	90.74	8,234	0.006467	1.260	24.92	40.12	794	0.05053
12	80.81	6,530	0.005129	1.588	19.77	50.59	629.6	0.08035
13	71.96	5,178	0.004067	2.003	15.68	63.80	499.3	0.1278
14	64.08	4,107	0.003225	2.525	12.43	80.44	396.0	0.2032
15	57.07	3,257	0.002558	3.184	9.858	101.4	314.0	0.3230
16	50.82	2,583	0.002028	4.016	7.818	127.9	249.0	0.5136
17	45.26	2,048	0.001609	5.064	6.200	161.3	197.5	0.8167
18	40.30	1,624	0.001276	6.385	4.917	203.4	156.6	1.299
19	35.89	1,288	0.001012	8.051	3.899	256.5	124.2	2.065
20	31.96	1,022	0.0008023	10.15	3.092	323.4	98.50	3.283
21	28.46	810.1	0.0006363	12.80	2.452	407.8	78.11	5.221
22	25.35	642.4	0.0005046	16.14	1.945	514.2	61.95	8.301
23	22.57	509.5	0.0004002	20.36	1.542	648.4	49.13	13.20
24	20.10	404.0	0.0003173	25.67	1.223	817.7	38.96	20.99
25	17.90	320.4	0.0002517	32.37	0.9699	1,031.0	30.90	33.37
26	15.94	254.1	0.0001996	40.81	0.7692	1,300	24.50	53.06
27	14.20	201.5	0.0001583	51.47	0.6100	1,639	19.43	84.37
28	12.64	159.8	0.0001255	64.90	0.4837	2,067	15.41	134.2
29	11.26	126.7	0.00009953	81.83	0.3836	2,607	12.22	213.3
30	10.03	100.5	0.00007894	103.2	0.3042	3,287	9.691	339.2
31	8.928	79.70	0.00006260	130.1	0.2413	4,145	7.685	539.3
32	7.950	63.21	0.00004964	164.1	0.1913	5,227	6.095	857.6
33	7.080	50.13	0.00003937	206.9	0.1517	6,591	4.833	1,364
34	6.305	39.75	0.00003122	260.9	0.1203	8,310	3.833	2,168
35	5.615	31.52	0.00002476	329.0	0.09542	10,480	3.040	3,448
36	5.000	25.00	0.00001964	414.8	0.07568	13,210	2.411	5,482
37	4.453	19.83	0.00001557	523.1	0.06001	16,660	1.912	8,717
38	3.965	15.72	0.00001235	659.6	0.04759	21,010	1.516	13,860
39	3.531	12.47	0.000009793	831.8	0.03774	26,500	1.202	22,040
40	3.145	9.888	0.000007766	1049.0	0.02993	33,410	0.9534	35,040

* From "Reference Data for Radio Engineers," International Telephone & Telegraph Corporation.

† Temperature coefficient of resistance: The resistance of a conductor at temperature t in degrees centigrade is given by

$$R = R_{20}[1 + \alpha_{20}(T - 20)]$$

where R_{20} is the resistance at 20°C and α_{20} is the temperature coefficient of resistance at 20°C. For copper, $\alpha_{20} = 0.00393$. That is, the resistance of a copper conductor increases approximately 0.4 per cent per degree centigrade rise in temperature.

TABLE A-7. ALLOWABLE CURRENT-CARRYING CAPACITY OF RUBBER- OR THERMOPLASTIC-COVERED WIRE

<i>Wire size, AWG</i>	<i>Current, amp</i>	<i>Wire size, AWG</i>	<i>Current amp</i>
20	3	4	70
18	5	3	80
16	7	2	95
14	15	1	110
12	20	0	125
10	30	00	145
8	40	000	165
6	55	0000	195

These values are for not more than three conductors in a raceway or cable and for a maximum temperature of 30°C or 86°F. Derating factors for higher temperatures are

°C	°F	Factor
40	104	0.82
45	113	0.71
50	122	0.58
55	131	0.41


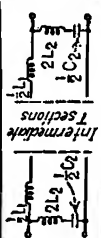
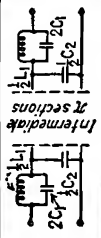

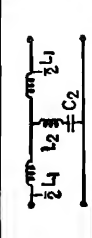
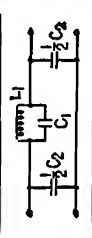

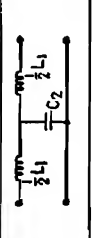
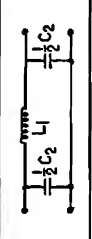
TABLE A-8a. DESIGN OF LOW-PASS SECTIONS

Fundamental Relations

R = load resistance f_2 = cut-off frequency f_∞ = a frequency of very high attenuation
 (highest frequency transmitted)

$$L_k = \frac{R}{\pi f_2} \qquad C_k = \frac{1}{\pi f_2 R} \qquad m = \sqrt{1 - \left(\frac{f_2}{f_\infty}\right)^2}$$

Design of Sections

Type	Attenuation characteristic	A. Filters having T intermediate sections		B. Filters having π intermediate sections	
		Configuration	Formulas	Configuration	Formulas
End (m of approximately 0.6)			$L_1 = mL_k$ $L_2 = \frac{1 - m^2}{4m} L_k$ $C_1 = mC_k$		$L_1 = mL_k$ $C_1 = \frac{1 - m^2}{4m} C_k$ $C_2 = mC_k$
I			$L_1 = mL_k$ $L_2 = \frac{1 - m^2}{4m} L_k$ $C_1 = mC_k$		$L_1 = mL_k$ $C_1 = \frac{1 - m^2}{4m} C_k$ $C_2 = mC_k$
II ($f_\infty = \infty$)			$L_1 = L_k$ $C_1 = C_k$		$L_1 = L_k$ $C_1 = C_k$

Source: F. E. Terman, "Radio Engineers' Handbook," McGraw-Hill Book Company, Inc., New York, 1943.

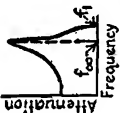
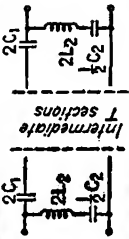
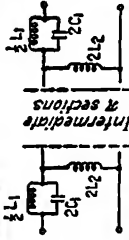
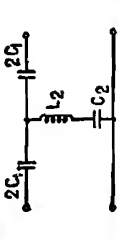
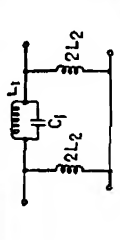
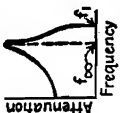
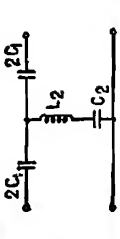
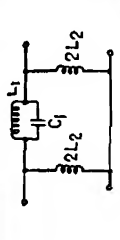

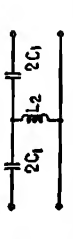
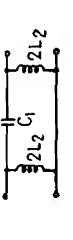
TABLE A-8b. DESIGN OF HIGH-PASS SECTIONS

Fundamental Relations

R_1 = load resistance f_1 = cut-off frequency f_∞ = a frequency of very high attenuation
 (lowest frequency transmitted)

$$L_b = \frac{R}{4\pi f_1} \qquad C_b = \frac{1}{4\pi f_1 R} \qquad m = \sqrt{1 - \left(\frac{f_\infty}{f_1}\right)^2}$$

Design of Sections

Type	Attenuation characteristic	A. Filters having T intermediate sections		B. Filters having π intermediate sections	
		Configuration	Formulas	Configuration	Formulas
End (m of approx- imately 0.6)			$C_1 = \frac{C_b}{m}$ $C_2 = \frac{4m}{1 - m^2} C_b$ $L_2 = \frac{L_b}{m}$		$L_1 = \frac{4m}{1 - m^2} L_b$ $C_1 = \frac{C_b}{m}$ $L_2 = \frac{L_b}{m}$
			$C_1 = \frac{C_b}{m}$ $C_2 = \frac{4m}{1 - m^2} C_b$ $L_2 = \frac{L_b}{m}$		$L_1 = \frac{4m}{1 - m^2} L_b$ $C_1 = \frac{C_b}{m}$ $L_2 = \frac{L_b}{m}$
I			$C_1 = \frac{C_b}{m}$ $C_2 = \frac{4m}{1 - m^2} C_b$ $L_2 = \frac{L_b}{m}$		$L_1 = \frac{4m}{1 - m^2} L_b$ $C_1 = \frac{C_b}{m}$ $L_2 = \frac{L_b}{m}$
II $f_\infty = 0$			$C_1 = \frac{C_b}{m}$ $L_2 = \frac{L_b}{m}$		$C_1 = \frac{C_b}{m}$ $L_2 = \frac{L_b}{m}$

SOURCE: F. E. Terman, "Radio Engineers' Handbook," McGraw-Hill Book Company, Inc., New York, 1943.

TABLE A-8c. DESIGN OF BAND-PASS SECTIONS

Fundamental Relations

 R = load resistance f_1 = lower frequency limit of pass band f_2 = higher frequency limit of pass band f_{∞} = a frequency of very high attenuation in low-frequency attenuating band f_{∞} = a frequency of very high attenuation in high-frequency attenuating band

$$L_{1k} = \frac{R}{\pi(f_2 - f_1)}$$

$$L_{2k} = \frac{(f_2 - f_1)R}{4\pi f_1 f_2}$$

$$C_{1k} = \frac{f_2 - f_1}{4\pi f_1 f_2 R}$$

$$C_{2k} = \frac{1}{\pi(f_2 - f_1)R}$$

Design of Sections

Type	Attenuation characteristic	A. Filters having T intermediate sections	B. Filters having π intermediate sections	Notation for both T and π sections
End ($m_1 = m_2$ = approximately 0.6)		Configuration: Formulas: $L_1 = m_1 L_{1k}$, $L_2 = a L_{1k}$, $L_1' = c L_{1k}$, $C_1 = \frac{C_{1k}}{b}$, $C_1' = \frac{C_{1k}}{d}$	Configuration: Formulas: $L_1 = \frac{L_{1k}}{b}$, $L_2 = \frac{L_{1k}}{a}$, $C_1' = m_1 C_{1k}$, $C_2 = m_1 C_{1k}$	$g = \sqrt{\left(1 - \frac{f_{1\infty}^2}{f^2}\right) \left(1 - \frac{f_{2\infty}^2}{f^2}\right)}$ $h = \sqrt{\left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right) \left(1 - \frac{f^2}{f_{2\infty}^2}\right)}$ $m_1 = \frac{f_{1\infty}^2}{f_{2\infty}^2} + h$, $m_2 = \frac{1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}}{1 - \frac{f^2}{f_{2\infty}^2}}$ $a = \frac{4\pi f_1 f_2}{(1 - m_1^2) f_{1\infty}^2} \left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right) = \frac{4\pi f_1 f_2}{(1 - m_1^2) f_{1\infty}^2} \left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right)$ $b = \frac{4\pi f_1 f_2}{(1 - m_2^2) f_{2\infty}^2} \left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right) = \frac{4\pi f_1 f_2}{(1 - m_2^2) f_{2\infty}^2} \left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right)$ $d = \frac{4\pi f_1 f_2}{(1 - m_1^2) f_{1\infty}^2} \left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right) = \frac{4\pi f_1 f_2}{(1 - m_1^2) f_{1\infty}^2} \left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right)$ when ($m_1 = m_2$), $g = h$, $a = d$, $b = c$, $f_{1\infty} = \frac{f_1 f_2}{f_{\infty}}$, $m_1 = m_2 = \frac{h}{1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}}$ and $f_{1\infty}^2 = \frac{f_1^2 + f_2^2 - 2m^2 f_1 f_2}{2(1 - m^2)} + \left[\left(\frac{f_1^2 + f_2^2 - 2m^2 f_1 f_2}{2(1 - m^2)} \right)^2 - f_1^2 f_2^2 \right]^{1/2}$
I		Configuration: Formulas: $L_1 = m_1 L_{1k}$, $L_2 = a L_{1k}$, $L_1' = c L_{1k}$, $C_1 = \frac{C_{1k}}{b}$, $C_1' = \frac{C_{1k}}{d}$	Configuration: Formulas: $L_1 = \frac{L_{1k}}{b}$, $L_2 = \frac{L_{1k}}{a}$, $C_1' = m_1 C_{1k}$, $C_2 = m_1 C_{1k}$	$g = \sqrt{\left(1 - \frac{f_{1\infty}^2}{f^2}\right) \left(1 - \frac{f_{2\infty}^2}{f^2}\right)}$ $h = \sqrt{\left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right) \left(1 - \frac{f^2}{f_{2\infty}^2}\right)}$ $m_1 = \frac{f_{1\infty}^2}{f_{2\infty}^2} + h$, $m_2 = \frac{1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}}{1 - \frac{f^2}{f_{2\infty}^2}}$ $a = \frac{4\pi f_1 f_2}{(1 - m_1^2) f_{1\infty}^2} \left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right) = \frac{4\pi f_1 f_2}{(1 - m_1^2) f_{1\infty}^2} \left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right)$ $b = \frac{4\pi f_1 f_2}{(1 - m_2^2) f_{2\infty}^2} \left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right) = \frac{4\pi f_1 f_2}{(1 - m_2^2) f_{2\infty}^2} \left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right)$ $d = \frac{4\pi f_1 f_2}{(1 - m_1^2) f_{1\infty}^2} \left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right) = \frac{4\pi f_1 f_2}{(1 - m_1^2) f_{1\infty}^2} \left(1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}\right)$ when ($m_1 = m_2$), $g = h$, $a = d$, $b = c$, $f_{1\infty} = \frac{f_1 f_2}{f_{\infty}}$, $m_1 = m_2 = \frac{h}{1 - \frac{f_{1\infty}^2}{f_{2\infty}^2}}$ and $f_{1\infty}^2 = \frac{f_1^2 + f_2^2 - 2m^2 f_1 f_2}{2(1 - m^2)} + \left[\left(\frac{f_1^2 + f_2^2 - 2m^2 f_1 f_2}{2(1 - m^2)} \right)^2 - f_1^2 f_2^2 \right]^{1/2}$
II $f_{1\infty} = 0$ $f_{2\infty} = f_2$		Configuration: Formulas: $L_1 = \frac{f_1 R}{\pi(f_2(f_2 - f_1))}$, $L_2 = \frac{f_1 R}{\pi(f_1(f_2 - f_1))}$, $C_1 = \frac{C_{1k}}{b}$	Configuration: Formulas: $C_1 = \frac{f_1 + f_2}{4\pi f_1 f_2 R}$, $C_2 = \frac{f_1}{\pi f_2(f_2 - f_1)R}$, $L_2 = \frac{L_{1k}}{a}$	

III $f_{\infty} = f_1$ $f_{\infty} = \infty$			$L_1 = L_{12}$ $C_1' = \frac{1}{\pi(f_1 + f_2)R}$ $C_1 = \frac{f_2 - f_1}{4\pi f_2 R}$		$L_1' = \frac{R}{\pi(f_1 + f_2)}$ $L_2 = \frac{(f_2 - f_1)R}{4\pi f_2^2}$ $C_1 = C_{12}$	$L_1 = L_{12}$ $L_2 = L_{22}$ $C_1 = C_{12}$ $C_2 = C_{22}$	$L_1' = \frac{R}{\pi(f_1 + f_2)}$ $L_2 = \frac{(f_2 - f_1)R}{4\pi f_2^2}$ $C_1 = C_{12}$
IV $f_{\infty} = 0$ $f_{\infty} = \infty$			$L_1 = L_{12}$ $L_2 = L_{22}$ $C_1 = C_{12}$ $C_2 = C_{22}$		$L_1 = L_{12}$ $L_2 = L_{22}$ $C_1 = C_{12}$ $C_2 = C_{22}$	$L_1 = L_{12}$ $L_2 = L_{22}$ $C_1 = C_{12}$ $C_2 = C_{22}$	$L_1 = L_{12}$ $L_2 = L_{22}$ $C_1 = C_{12}$ $C_2 = C_{22}$
V $f_{\infty} = f_2$			$L_1 = m_1 L_{12}$ $L_2 = \frac{1}{(1 - m_1^2)} L_{22}$ $C_1 = \frac{m_1}{m_2}$ $C_2 = \frac{4m_2}{1 - m_1^2} C_{12}$ See notation for m_1 and m_2		$L_1' = \frac{4m_2}{1 - m_1^2} L_{22}$ $L_2 = \frac{m_1}{m_2}$ $C_1 = (1 - m_1^2) C_{22}$ $C_2 = m_1 C_{12}$ See notation for m_1 and m_2	$m_1 = \frac{f_1}{f_2} m_2$ $m_2 = \sqrt{\frac{1 - \frac{f_1^2}{f_2^2}}{1 - \frac{f_1^2}{f_2^2}}}$	
VI $f_{\infty} = f_1$			Same circuit as above for Type V		Same formulas as above for Type V See notation for m_1 and m_2	$m_1 = \sqrt{\frac{f_1^2}{1 - \frac{f_1^2}{f_2^2}}}$ $m_2 = \frac{f_1}{f_2} m_1$	
VII $f_{\infty} = 0$			$L_1 = m_1 L_{12}$ $L_2 = a L_{22}$ $L_1' = \frac{1}{(1 - m_1^2)} L_{22}$ $C_1 = C_{12}$ $C_1' = \frac{h}{a} C_{12}$		$L_1' = \frac{h}{a} L_{22}$ $L_2 = L_{22}$ $C_1 = \frac{a}{C_{12}}$ $C_2 = m_1 C_{12}$ $C_1' = \frac{1}{(1 - m_1^2)} C_{12}$	$h = \sqrt{\left(1 - \frac{f_1^2}{f_2^2}\right) \left(1 - \frac{f_1^2}{f_2^2}\right)}$ $m_1 = \frac{f_1}{f_2} + h$ $a = \frac{f_2^2}{(1 - m_1^2) f_2^2}$	
VIII $f_{\infty} = \infty$			$L_1 = L_{12}$ $L_2 = \frac{d}{\theta} L_{22}$ $C_1 = \frac{C_{12}}{m_1}$ $C_2 = \frac{4\theta}{1 - m_1^2} C_{12}$ $C_1' = \frac{C_{12}}{\theta}$		$L_1 = \frac{4\theta}{1 - m_1^2} L_{22}$ $L_2 = \frac{m_1}{\theta} L_{22}$ $L_1' = \frac{d}{\theta}$ $C_1 = \frac{d}{C_{12}}$ $C_2 = C_{12}$	$\theta = \sqrt{\left(1 - \frac{f_1^2}{f_2^2}\right) \left(1 - \frac{f_1^2}{f_2^2}\right)}$ $m_1 = \theta + \frac{f_1^2}{f_2^2}$ $d = \frac{f_2^2}{(1 - m_1^2) f_2^2}$	

SOURCE: F. E. Terman, "Radio Engineers' Handbook," McGraw-Hill Book Company, Inc., New York, 1943.

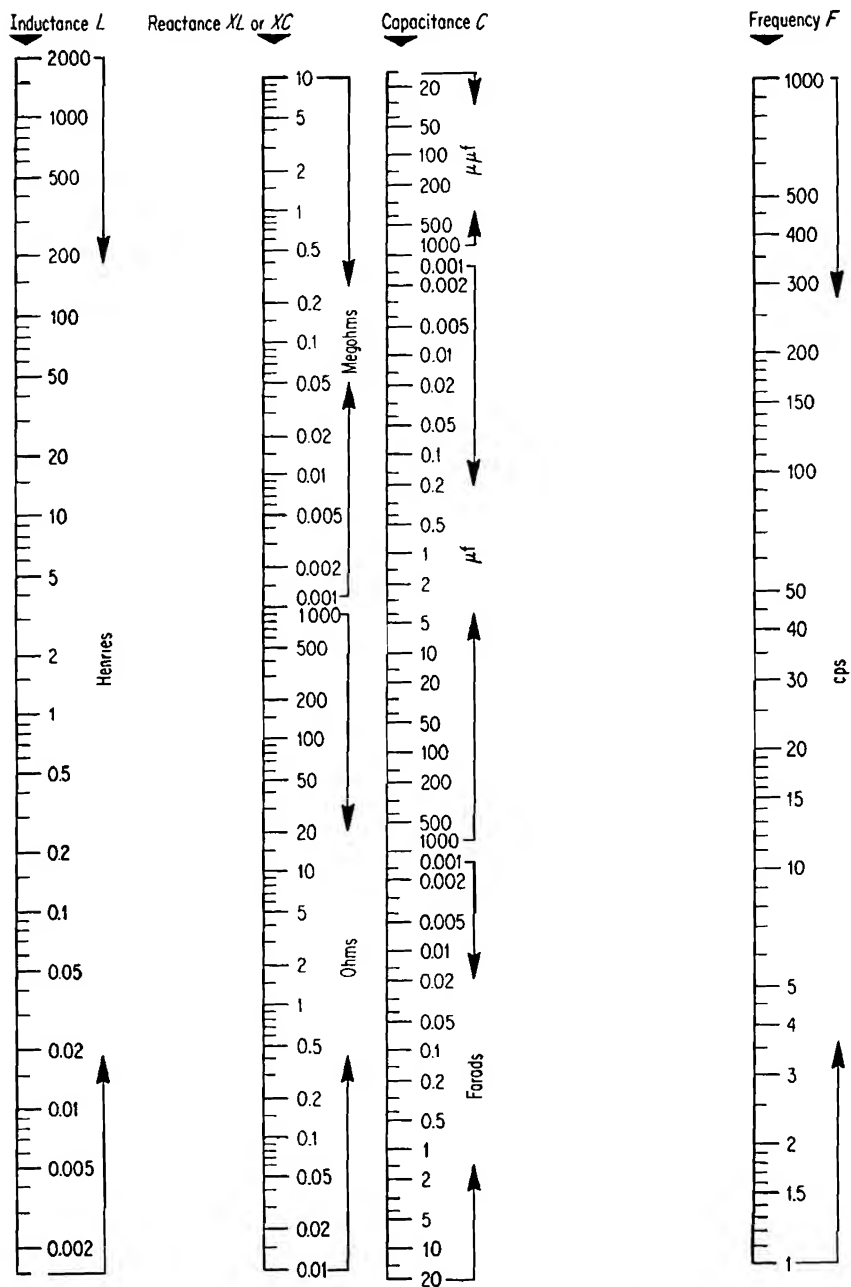


FIG. A-1a. Reactance chart covering 1 to 1,000 cps. (From "Reference Data for Radio Engineers," ITT.)

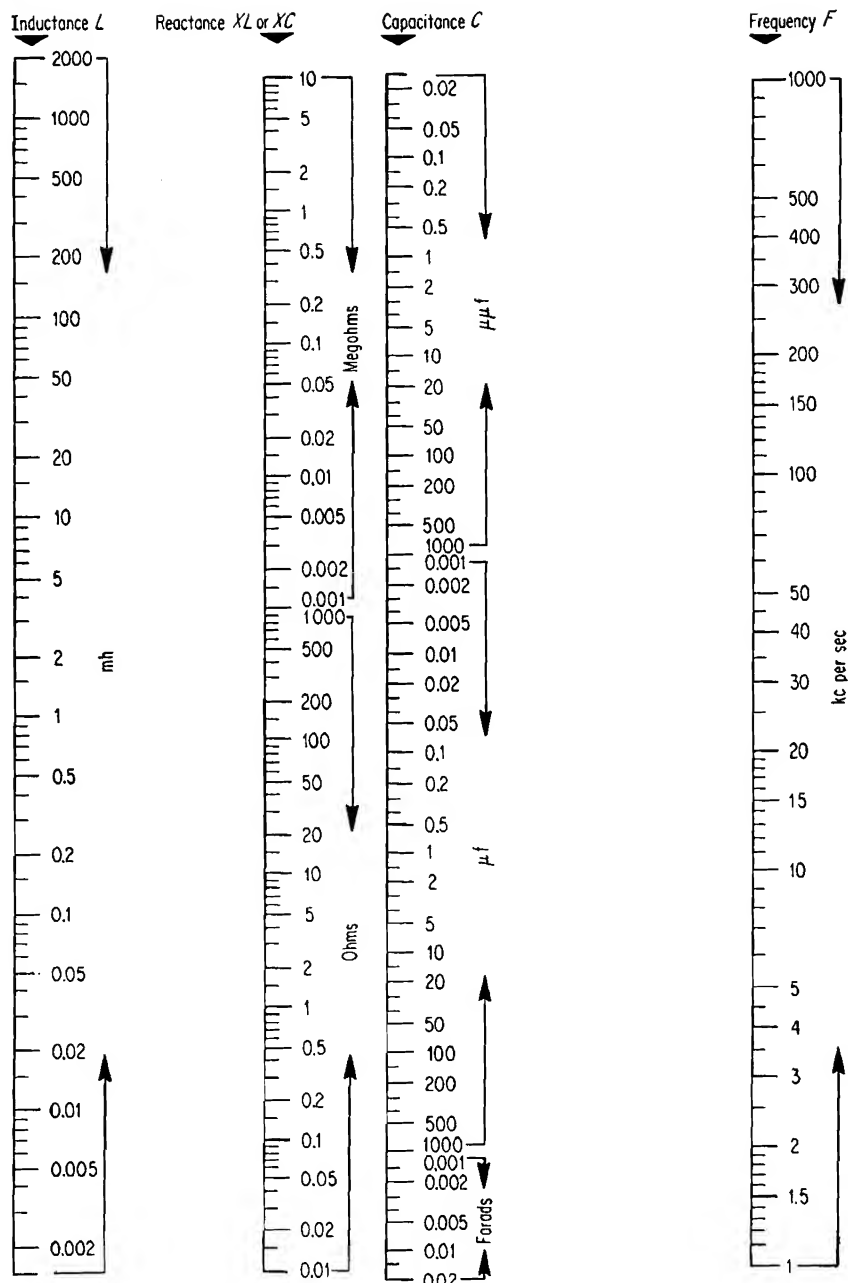


FIG. A-1b. Reactance chart covering 1 to 1,000 kc. (From "Reference Data for Radio Engineers," ITT.)

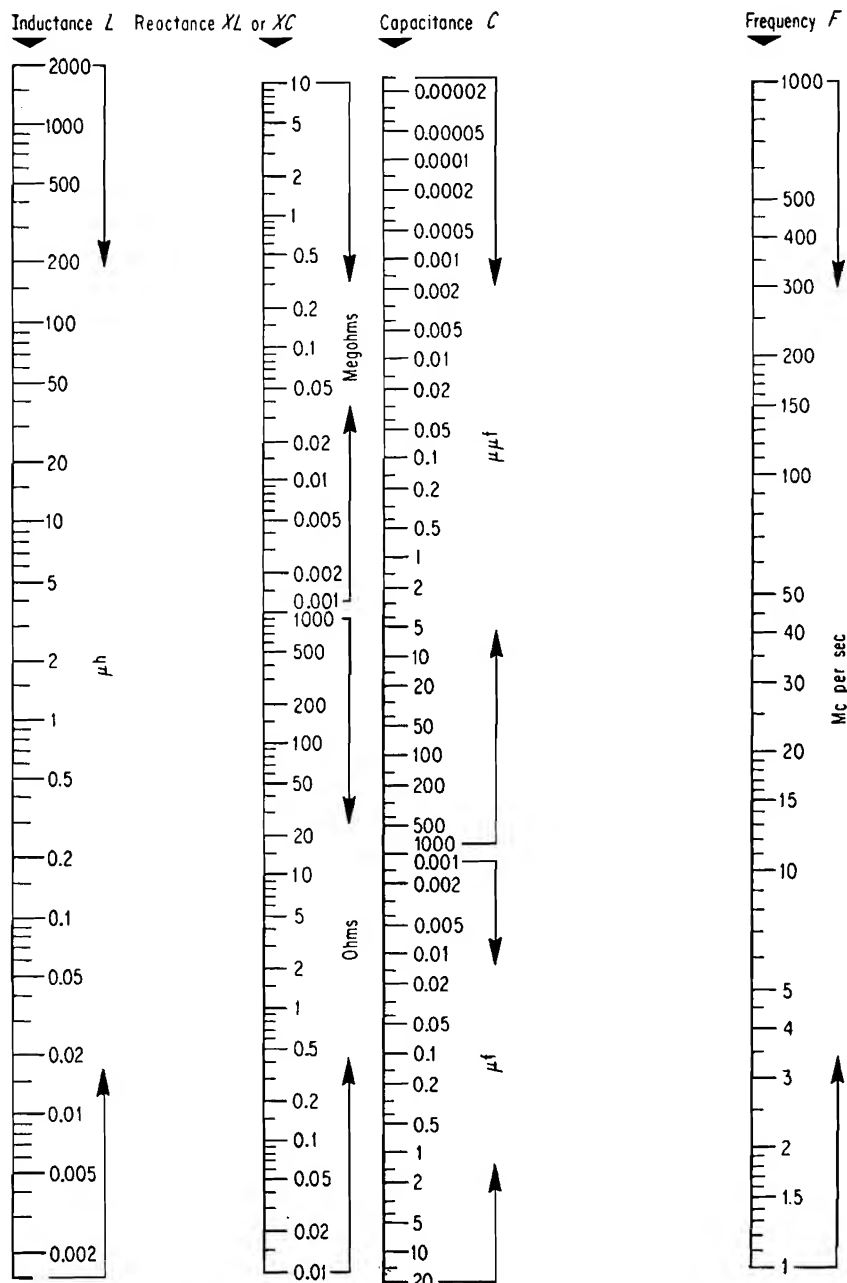


FIG. A-1c. Reactance chart covering 1 to 1,000 Mc. (From "Reference Data for Radio Engineers," ITT.)

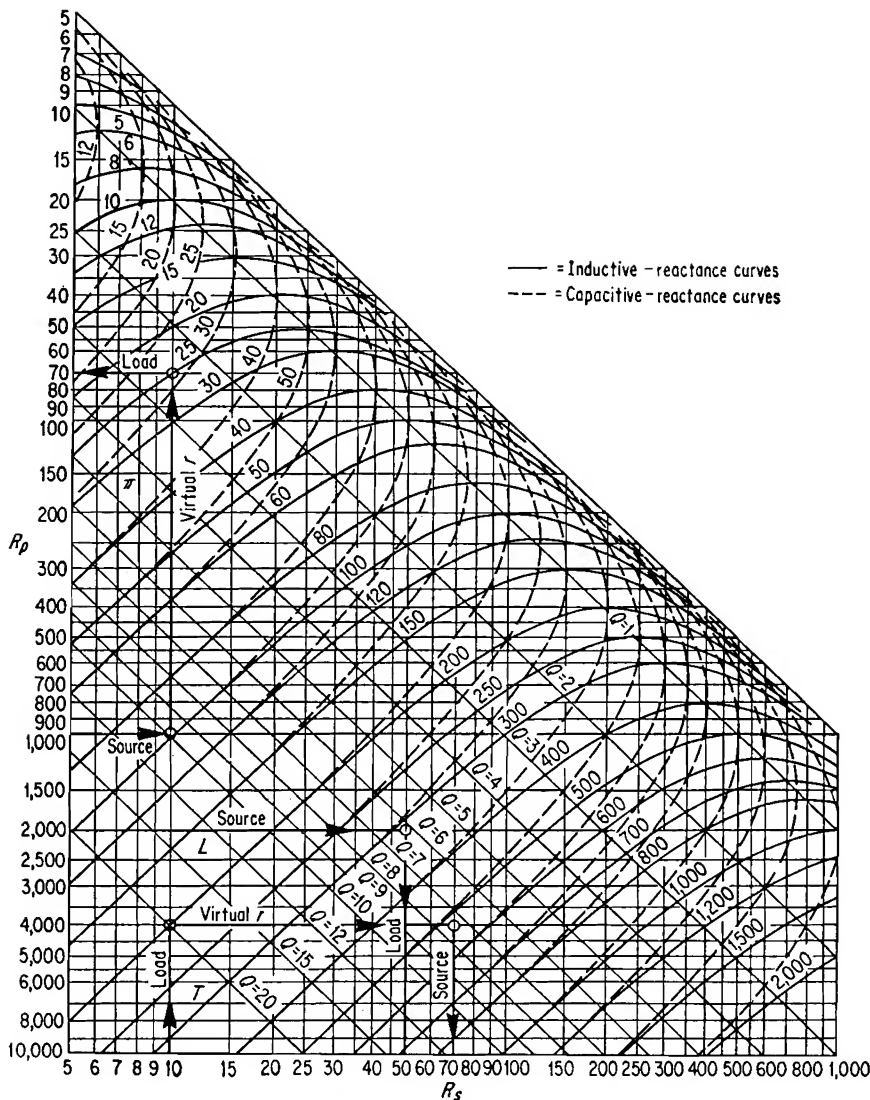
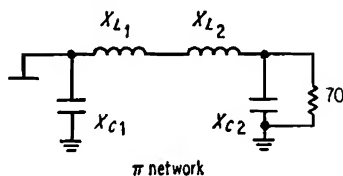
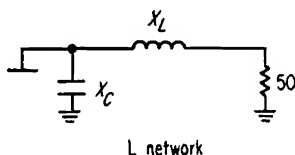


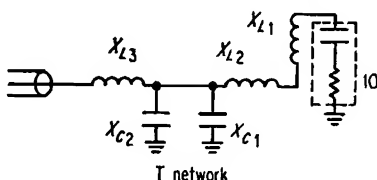
FIG. A-2. Universal curves for π , L , and T networks. These universal curves are derived from the equations given in Sec. 5-7. Examples are given in the text.

How to Use Universal Curves for π , L, and T Networks

π Network. A vacuum tube with a required a-c plate load impedance of 1,000 ohms is to be matched to a 70-ohm transmission line. Enter the chart at $R_P = 1,000$ and move horizontally until the dashed line corresponding to the selected value of Q is intersected. Say the selected value of Q is 10. This point gives the value of capacitive reactance to be placed across the tube which is 100 ohms. Then move up the chart until the horizontal line corresponding to $R_P = 70$ is intersected. This point gives the value of capacitive reactance to be placed across the transmission line and this is 28.5 ohms. This same point gives the value of inductive reactance to be placed in series with the line which is 24.5 ohms. The total value of inductive reactance in the network will be $X_{L1} + X_{L2} = 124.5$ ohms.



L Network. A vacuum tube with a required a-c plate load impedance of 2,000 ohms is to be matched to a 50-ohm transmission line. Enter the chart at $R_P = 2,000$ and move over until the vertical line corresponding to $R_S = 50$ is reached. This point gives the value of $Q = 6.3$, and $X_C = X_L = 317$ ohms.



T Network. An antenna with a resistance of 10 ohms and capacitive reactance of 100 ohms is to be matched to a 70-ohm transmission line. The capacitive reactance must first be balanced out with a 100-ohm inductive reactance X_{L1} . Enter the chart at $R_S = 10$ and move up until the maximum allowable value of Q is reached. Say this is 20. This point gives the inductive reactance to be connected in series with the 100-ohm antenna loading inductance. The value is 200 ohms. This point also gives the shunt capacitive reactance which is 200 ohms. From this point move over until the vertical line corresponding to $R_S = 70$ ohms is intersected. This point gives the inductive reactance to be connected in series with the transmission line and this value is 533 ohms. This point also gives the shunt capacitive reactance which is also 533 ohms. The two capacitors in parallel will result in a total capacitive reactance of 145 ohms.

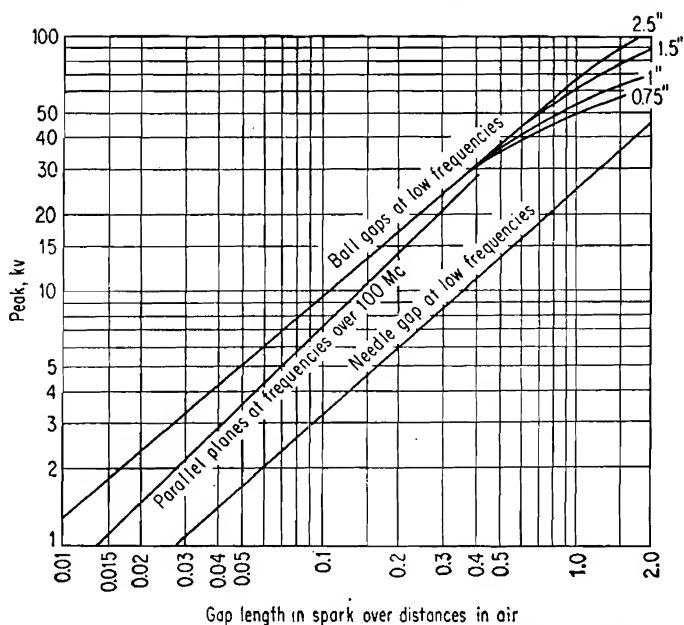


FIG. A-3. Spark-over distances in air. (From "Reference Data for Radio Engineers," ITT.)

INDEX

- Absorption wattmeter, 340
- Absorption wavemeter, 357
- A-c voltage regulators, 206
- Accelerator, linear, 329
- Accentuation, preemphasis, 186
- Across-the-line diagrams, 268
- Across-the-line starter, 245
- Admittance meter, 394
- Air cooling, tubes, 270
- Air pressure, measurement, 276
- Air temperature, tube outlet, 275
- Air velocity, measurement, 276
- Alarms, 250
- All-pass filters, 325
- Amplifier, amplatron, 105
 - bias, 84
 - class A, 38, 40
 - class AB, 38
 - class B, 38, 43
 - class C, 38, 46
 - conduction angle, 39
 - efficiency, 39
 - grounded-cathode, 38
 - grounded-grid, 38
 - grounded-plate, 38
 - klystron, 95
 - linear, 72
 - distortion in, 76
 - r-f, 152
 - linearity measurement, 79
 - negative-grid, 95
 - neutralization, 64
 - output load resistance, 60
 - parasitic oscillations, 62
 - platinotron, 104
 - stabilization, 61
 - television, linearity, 391
 - traveling-wave tube, 100
 - tube performance computer, 47
 - tuned, 380
 - wideband, 80
- Amplatron, 104
- Amplitude modulation (*see* Modulation, amplitude)
- Analyzer, spectrum, 134, 195, 358
- Angle modulation, 181
- Anodes, 93
 - cooling methods, 93
 - external, 94
 - forced-air-cooled, 94
 - liquid-cooled, 95
 - modulating, 150
- Antenna, dummy, 339
- Anti-TR (ATR) switch, 312
- Arcback in rectifiers, 220
- Armstrong f-m system, 188
- ARQ system, 314
- Artificial respiration, 418
- Askarel, 229
- Astable multivibrator, 196
- ATR switch, 312
- Attenuation, after deemphasis network, 186
 - r-f cables, 342
- Attenuator values, 435
- Audio oscillator, 368
- Audio power measurement, 335
- Automatic-error-correcting (ARQ) system, 314
- Automatic frequency control, 9
- Automatic recycling, 251
- Automatic tuning, 259
- Avigation, 309
- Back porch, 322, 384
- Backward-wave oscillator, 102
- Backwave, 137
- Balanced modulator, 29, 141, 162
 - lattice or ring, 164
 - push-pull, 164
 - push-push, 165

- Balanced modulator, series-bridge, 163
 - shunt-bridge, 163
- Balanced-to-unbalanced coupling networks, 125
- Balun, 125
- Bands, frequency, 4, 311
- Bandwidth, double-tuned circuit, 113
 - f-m signal, 184
 - single-tuned circuit, 110
 - television signal, 321
 - various types of transmission, 3
- Barretter, 341
- Baud, 3
- BC-221 frequency meter, 356
- Bessel functions, 183
- Bev, 425
- Binary coding, 193
- Biphase keying, 314
- Bistable multivibrator, 28, 196
- Bit, 3, 194
- Black negative TV signal, 383
- Bleeder resistance, 228, 420
- Blocked-grid keying, 138
- Blocking oscillator, 28, 195
- Blower, calculations, 272-276
 - high-altitude operation, 273
 - high-temperature operation, 275
- Bolometer, 341
- Bombarder cathode, 92
- Break-in, 137, 313
- Breather, 426
- Bridge, bolometer, 341
 - impedance, 393
- Bridge-stabilized oscillator, 20
- Broadcast, multiplex transmission, 320
 - repeaters and translators, 328
 - standard frequency, 347
 - transmitters, 319
- Buck-boost transformer, 214
- Buffer capacitor, 240
- Bunching, optimum, 99
- Burst, color, 326

- Cabinet radiation, 406
- Calibration, amplitude modulation
 - meter, 362
- Call signal prefixes, international, 431
 - United States, 433
- Capacity, interrupting, 422
- Carbon tetrachloride, 426
- Carcinotron, 102
- Cardiac massage, 419
- Carrier, 5
- Carrier disappearance in f-m, 183, 364
- Carrier shift, 142, 362
- Carrier suppression, 152
- Cascode amplifier, 151
- Cathode, 89-92
 - bombarder, 92
 - filamentary, 92
 - indirect-heated, 92
 - oxide-coated, 91
 - thoriated tungsten, 91
 - tungsten, 90
- Cathode follower, 60, 191
- Cathode modulation, 149
- CCIR (International Radio Consultative Committee), 3
 - frequency range designations, 4
 - method of measuring peak-envelope power, 339
- Cesium, atomic oscillation, 348
- Chart, reactance, 445-447
 - Smith, 396
- Chemical hazards, 426
- Chirp, 137
- Chlorinated biphenyl, 426
- Chokes, 228-231
 - filter, 228
 - swinging, 227
- Chopper, 383
- Chrominance signal, 326
- Circuit breaker, magnetic, 263
 - relay trip, 264
 - series trip, 264
 - shunt trip, 264
 - thermal, 263
 - undervoltage trip, 264
- Circulating current in tuned circuit, 110
- Circulator, ferrite, 306
- Clamp, keyed, 324
- Clamp-tube modulator, 152
- Clamped signal, 322
- Clicks, key, 137
- Clipping, biased diode, 196
 - cutoff, 197
 - grid limiting, 197
 - speech, 316
- Coaxial hybrid ring, 298
- Coefficient, reflection, 403
 - temperature, 11

- Color burst, 326
- Color subcarrier, 325
 - amplitudes, 326
 - phase angles, 326
- Color transmission, 325
- Colpitts oscillator, 12
- Combining transmitter outputs, 298
- Companzor, 7
- Compatible single sideband, 319
- Conduction angle, 39
- Conservation, spectrum, 4, 157, 405
- Constant-duty-cycle modulation, 310
- Contactors, 244
- Continuous wave (c-w), 136
 - interrupted, 318
 - modulated, 318
- Control-grid modulation, 146
- Controlled-carrier modulation, 151
- Conversion, p-m to f-m, 184
- Copper wire table, 438
- Corona, 231
- Correction factor for ethylene glycol, 341
- Coupling circuits, 108-133
 - balanced-to-unbalanced, 125
 - double-tuned, 81, 112
 - L network, 117
 - link, 122
 - parallel-tuned, 20, 108
 - pi network, 118
 - pi-L network, 121
 - series-tuned, 20, 111
 - T network, 120
- Cross, pulse, 387
- Cross filter, high-low, 392
- Crossover, spurious frequency, 28
- Crowbar, electronic, 265
- Crystal, activity, 22
 - aging, 23
 - axes, 13
 - cuts, AT, 14, 19, 22
 - GT, 14, 23
 - rotated X-cut, 14
 - rotated Y-cut, 14
 - drive level, 21
 - equivalent circuit, 19
 - frequency correlation, 16
 - MIL designations, 16-18
 - overtones, 13
 - series resistance, 22
 - temperature coefficient, 16
- Crystal calibrator, 354
- Crystal filter, 166
- Crystal holder, 14, 15
- Crystal oscillator (*see* Oscillator)
- Crystal oven, 16
- Current-carrying capacity of wire, 439
- Cytac, 310
- D-c regulators, 236
- D-c transmission, 321
- Decca, 310
- Decibel reference levels, 437
- Decibel table, 436
- Decommutator, 315
- Deemphasis, 7, 185
- Delay, envelope, 115, 327, 389
- Delay equalizers, 328
- Delay multivibrator, 196
- Delta connection, 335
- Delta modulation, 201
- Demodulation, 6
- Density modulation, 98
- Density of r-f radiation, 422
- Detection, phase-lock, 6
- Deviation, frequency, 182, 352
 - maximum for mobile use, 317
 - phase, 182
 - ratios for carrier disappearance, 183
- Diagonal clipping, 327
- Diathermy, 329
- Dielectric heating, 329
- Differential gain, 391
- Differential phase, 327
- Differentiation of square waves, 195
- Diode modulator, 139
- Diplexer, 294-300, 312
- Direct f-m, 189
- Directional couplers, 283-290, 398
 - loop type, 286
 - resistance type, 284
 - single-hole, 288
 - two-element, 289
- Display, TV waveform, 382
- Distortion, harmonic, 369
 - intermodulation, 369
 - telegraph, 372
- DMET (distance measuring element of Tacan), 310
- Door interlocks, 419
- Doppler effect, radar, 312
 - on single-sideband transmission, 318

- Doppler effect, on standard-frequency transmissions, 346
- Double-sideband amplitude modulation, 136
 - reduced carrier, 153
- Double-stub tuner, 131, 396
- Double-tuned circuit, 81, 112
- Downward modulation, 141
- Drift-canceled oscillator, 30
- Drift tube in klystrons, 95
- Drive level, crystal, 21
- Driver, modulator, 145
 - r-f, 142
- Dummy antenna, 339
- Dummy loads, 290-292
- Duplexer, 312
- Duty cycle, 195
- Dynamotor, 238
- Efficiency, amplifier, 39
 - class A, 41
 - class B, 44
 - class C, 46
 - modulation, 137
 - r-f matching network, 128
- Electric shock, 417-419
- Electron-coupled oscillator, 12, 24
- Electron gun, 95
- Electron volt, 425
- Electronic crowbar, 265
- Electronic frequency counter, 353
- Electronic weatherstripping, 406
- Emission classifications, 2
- Envelope delay, 115, 327, 389
- Equalizers, delay, 328
- Ethylene glycol, specific heat and gravity, 341
- Evaporative cooling, 93
- Exalted carrier, 152
- Exposures, X-ray, 425
- Eye cataracts, 422
- Facsimile, 320
- Factor, conversion, words per minute to bauds or cps, 3
 - correction, for ethylene glycol in calorimeters, 341
 - power, measurement, 334
- Faraday rotation in ferrites, 305
- Fault location in transmission line, 400
- Federal Communications Commission, 1, 7
- Feedback, negative, 155
- Ferrites, 305
- Field strength, measurements, 405
 - of unlicensed transmitters, 1
- Filament, d-c supply, 235
 - multiphase, 235
 - starting, 232
- Filter, all-pass, 325
 - capacitor-input, 224
 - choke input, 225
 - crystal lattice, 168
 - design charts, 440-443
 - key-click, 138
 - mechanical, 169
 - phase-correcting, 325
 - power supply, 224
 - quartz crystal, 166
 - r-f, 292
 - ripple factor, 225
- Fin cooler, 94
- Flip-flop circuit, 196
- Flux, power, in front of antenna, 423
- Flywheel effect, 140
- Forced-air cooling, 270
- Forecasts, propagation, 345
- Frequency, intercarrier, 352
 - international distress, 318
 - spurious, 4, 28, 405
 - transmission-line cutoff, 406
 - waveguide cutoff, 401
- Frequency allocations, 3
- Frequency bands, 4, 311
- Frequency changing, 134
- Frequency control, automatic, 9
 - in f-m transmitters, 24
 - of microwave oscillators, 25
 - of reflex klystrons, 26
- Frequency conversion, 134
- Frequency deviation, 182, 352
 - maximum for mobile use, 317
- Frequency divider, 27
- Frequency division, multiplex, 315
- Frequency domain, 182
- Frequency measurements (*see* Measurements, frequency)
- Frequency-measuring stations, 349
- Frequency meters, 353
 - direct-reading, 356, 357
 - heterodyne, 355, 356

- Frequency mixer, 28, 173
 - crossovers in, 28, 30
- Frequency modulation (*see* Modulation, frequency)
- Frequency monitors, 351
- Frequency multiplier, 27, 82
- Frequency-range designations, 4, 311
- Frequency response, 375, 381
- Frequency-shift keying, 22, 137, 313-315
- Frequency spectrum, a-m, 135
 - f-m, 183
 - pulse, 195
- Frequency stability, crystal oscillator, 23
 - inductance-capacitance oscillator, 10
 - long-term, 9
 - medium-term, 9
 - mobile transmitters, 95
 - short-term, 9, 23
- Frequency standards, 344, 348
- Frequency stepping (FSK), 315
- Frequency synthesizer, 9, 27, 32, 33
- Frequency translation, 134, 172, 328
- Fresnel region, 423
- Front porch, 383
- FSK, 313
- Functions, Bessel, 183
- Fuse, types, 262
 - voltage rating, 422
- Gain, differential, 391
- Generator, pulse, 195
 - Thévenin, 379
- Genetic cell damage, 422
- Geomagnetic disturbances, 346
- Gray scale, 327
- Grid-dip meter, 358
- Grid pulsing, 198
- Grids, 93
- Grounding methods, 421
- Grounding sticks, 420
- Grounding switches, 420
- H-pad values, 435
- Half-power point, 110
- Hard-tube modulator, 197
- Harmonic, 5, 292, 405
- Harmonic distortion, 369
- Harmonic generators, 82
- Hartley oscillator, 12
- Hazards, chemical, 426
 - lightning, 421
 - radiation, 422-426
- Heat exchanger, liquid-to-air, 278
 - liquid-to-liquid, 278
- Heating, dielectric, 329
 - induction, 329
- Heterodyne frequency meter, 355
- High-altitude blower operation, 273
- High-low cross filter, 392
- Hybrid ring, 298
- Icw, 318
- Ideal detector curve, 375
- Idealized picture transmitter characteristic, 321
- Identification of transmitters, 2
- Impedance, bridge, 393
 - of parallel-tuned circuit, 109
 - reactive load, 122
- Impedance matching, of transmission lines, 129
 - of waveguide, 132
- Impedometer, 403
- In-band noise, 405
- Incandescent lamps, 250
- Incidental radiation device, 1
- Independent sideband (isb), 316
- Index, modulation, 182
- Indicators, 249
 - lamp life, 250
- Indirect f-m, 189
- Induction heating, 329
- Intelligibility of single-sideband speech, 318
- Intercarrier frequency, 352
- Interferometer, 359
- Interlock, door, 419
- Interlocking circuits, 247
- Intermodulation, 5
 - distortion, 141
 - measurement, 370, 406
 - in power amplifiers, 406
 - single-sideband, 339
- International call sign prefixes, 431
- International Frequency Registration Board, 3
- International Radio Consultative Committee (*see* CCIR)

- International Telecommunication Union (ITU), 3
- Interrupted continuous wave (icw), 318
- Ionization, air, 231
- IRE roll-off, 385
- Irreversible effects of radiation, 422
- Isb, 316
- Isolator, 305

- JJY, 347
- Junction transistor, 22

- Key-click filters, 138
- Key clicks, 137
- Keyed clamp, 324
- Keyer, multiplex, 315
- Keying, biphase, 316
 - blocked-grid, 138
 - clamp tube, 138
 - frequency-shift, 313
 - on-off, 137, 313
 - quadriphase, 316
 - transients, 138, 313
- Kilobits, 3
- Kineplex, 315
- Klystron amplifier, 95
 - amplitude modulation of, 150
 - drift tube, 95
 - frequency modulation of, 193
 - gain or bandwidth of, 95
 - reflex, 26

- L network, 117
 - universal curves, 448, 449
- Lamps, neon and incandescent, 250
- Licensing of transmitters, 1
- Lightning hazards, 421
- Lilt, 137
- Line-type pulser, 197
- Linear accelerator, 329
- Linear amplifier, 72, 76
 - r-f, 152
- Linear phase characteristic, 328
- Linearity, differential, 327
 - of TV amplifiers, 391
- Link coupling, 123
- Liquid cooling systems, 277
 - maintenance, 279
- Liquid insulation, 229
- Liquid-to-air heat exchanger, 278
- Liquid-to-liquid heat exchanger, 278
- Lissajous figures, 404
- Load, dummy, 290, 339
- Localizer transmitters, 309
- Location of faults in transmission line and waveguide, 400
- LOL, 347
- Long-lines effect, 306
- Long-term stability of crystal oscillators, 23
- Loran, 310
- Losses in air systems, 271

- Mach number, 318
- Magic T, 297
- Magnetic circuit breaker, 263
- Magnetostriction filters, 169
- Magnetron, 103
 - performance chart, 104
 - Rieke diagram, 104
 - rising sun structure, 104
 - strapped, 104
- Margin, printer, 372
- Mark signal, 3, 313
- Matching network (*see* Coupling circuits; Networks)
- Mcw, 318
- Measurements, 333-416
 - amplifier frequency response, 375
 - point-by-point, 379
 - sideband analyzer, 376
 - sweeping oscillator, 378
 - transient method, 382
 - audio power, 335
 - broad-band multiplex, 367
 - demodulated audio, 367
 - distortion, 369
 - field strength, 405
 - frequency, direct-beating method, 349
 - direct-reading meters, 356
 - electronic counter, 351
 - grid-dip meter, 358
 - interferometer, 359
 - paramagnetic resonance, 359
 - sliding harmonic method, 355
 - spectrum analyzer, 358
 - synthesizer, 32, 356
 - transfer oscillator method, 350
 - wavemeter, 357

- Measurements, harmonic, 405
 - harmonic distortion, 369
 - impedance, directional coupler, 403
 - slotted line, 396
 - in-band noise, 407
 - input peak-envelope power, 338
 - intermodulation, multichannel, 374
 - r-f, 406
 - intermodulation distortion, 369
 - intermodulation noise, 374
 - modulation, amplitude, 361
 - frequency, 363
 - noise, 370
 - noise-loading, 373
 - peak-envelope power, CCIR method, 339
 - two-tone test, 338
 - phase, with Lissajous figures, 404
 - power, primary input, 333
 - r-f output, 343
 - power factor, 334
 - printer margin, 372
 - pulse, 375
 - spurious output, 405
 - telegraph distortion, 372
 - television, color, 388
 - differential gain, 391
 - differential phase, 392
 - envelope delay, 389
 - peak power, 344
 - vertical-interval tests, 392
 - waveform, 382
 - time, 360
 - voltage-standing-wave ratio, 396, 403
 - X radiation, 426
 - Microwave oscillator, 25
 - frequency control, 26
 - MIL-C-3098C, 16
 - Military specification, quartz-crystal
 - units, 16
 - radiation levels, 422
 - Miller effect, 191
 - Mixer, balanced modulator, 28
 - frequency, 28, 173
 - crossover chart, 30
 - Mixing in power amplifiers, 405
 - Mobile transmitters, 317
 - Modulated continuous wave (mcw), 318
 - Modulating anode, 150
 - Modulation, angle, 181
 - amplitude, 134-180
 - Modulation, amplitude, double-sideband, 136
 - measurement, 361
 - single-sideband, 157
 - cathode, 149
 - choice of, 5
 - constant-duty-cycle, 310
 - constant-voltage, 146
 - control-grid, 146
 - controlled-carrier, 151
 - delta, 201
 - downward, 141
 - efficiency, 148
 - frequency, bandwidth used, 184
 - deemphasis, 185
 - direct generation, 189
 - indirect generation, 189
 - improvement in signal-to-noise ratio, 6
 - of klystrons, 193
 - measurement, 363
 - mechanical generation, 189
 - reactance-tube, 190
 - sidebands in, 183
 - threshold effect, 6
- general process, 5
- grid, 146
- Heising, 144
- Kahn, 319
- monitors, 362
- negative feedback, 155
- outphasing, 137
- percentage, 136, 362
- phase, generating methods, 186
- phase-to-amplitude, 154
- plate, 141
- pulse, 193
 - conversion to f-m, 184
- pulse-amplitude, 198
- pulse-code, 193, 200
- pulse-count, 193
- pulse-duration, 199
- pulse-frequency, 200
- pulse-position, 200
- pulse-width, 199
- of RC oscillators, 191
- screen-grid, 148
- suppressed-carrier, 152
- suppressor-grid, 148
- Modulation index, 182
- Modulator, balanced, 29, 141, 162-165

- Modulator, clamp-tube, 152
 - controlled multivibrator, 189
 - diode, 139
 - frequency, Armstrong, 188
 - controlled-diode, 192
 - Miller effect, 191
 - reactance-tube, 190
 - Serrasoid, 189
 - hard-tube, 197
 - klystron, 150
- Monitor, amplitude modulation, 362
 - frequency deviation, 351
 - frequency modulation, 364
 - phase, 403
- Monkey chatter, sideband, 319
- Monochrome television signal, 375
- Monostable multivibrator, 196
- Mouth-to-mouth technique, 419
- Multiburst, 378
- Multiplex, facsimile, 320
 - frequency-division, 315
 - keyers, 315
 - stereophonic, 320
 - time-division, 315
 - transmission, broadcast, 320
- Multivibrators, 28, 196
- Navigation, 309
- Negative feedback, 155
- Neon lamp, 250
- Networks, coupling, 108
 - L, 117, 448
 - π , 118, 448
 - T, 120, 448
 (See also Coupling circuits)
- Neutralization, 64
- Noise, in-band, 405
 - increase when sources combined, 374
 - loading test, 373
 - measurements, 370
 - objective level in United States toll circuits, 374
- Noise-improvement systems, 6
- Nonsynchronous vibrator power supply, 240
- Notch diplexer, 296
- O-pad values, 435
- Ohm's law, 336, 417
- Oil-filled transformers, 229
- Oscillation, parasitic, 62, 139, 405
- Oscillator, audio, 368
 - backward-wave, 102
 - blocking, 28, 195
 - bridge-stabilized, 20
 - carcinotron, 102
 - Clapp, 10
 - Colpitts, 12
 - crystal, Butler, 21
 - electron-coupled, 24
 - Miller, 20
 - Pierce, 20
 - transistor, 21
 - variable-frequency, 22
 - disciplined, 27
 - drift-canceled, 30
 - electron-coupled, 12
 - Hartley, 12
 - inductance-capacitance, 10
 - magnetron, 103
 - microwave, 25
 - phase-locked, 25, 314
 - platinotron, 105
 - reflex klystron, 99
 - transfer, 350, 366
 - tuned-plate tuned-grid, 12
 - tuning-fork, 27
 - variable-frequency, 22
- Oscillograph, 359
- Outphasing modulation, 154
- Ovens, crystal, 16
- Overload circuits, 260
- Overmodulation, 405
- Ozone, 231
- Panoramic adapter, 359
- Parallel-resonant circuit, 20, 108
- Parasitic oscillation, 62, 135, 405
- Peak-envelope power, 337
- Peaks, clipping of, 316
- PEP, 337
- Percentage of modulation, 136, 362
- Phantastron, 28, 197
- Phase, linear characteristic, 328
- Phase adjustment, power divider, 301
- Phase-to-amplitude modulation, 154
- Phase deviation, 182, 183
- Phase-locked oscillator, 6, 34, 314
- Phase modulation (see Modulation, phase)

- Phase monitor, 403
- Phase response, 324
- Phase-shift network, a-f, 170
 - r-f, 171
- Phase-shift Teletype, 314
- Phasitron, 189
- Photometric method of measuring
 - r-f power, 343
- Pi network, 118
 - universal curves, 448, 449
- Pierce oscillator, 20
- Piezoelectric effect, 12
- Plate modulation, 141
- Platinotron, 104
- Polarity of TV signal, 383
- Polymer fume fever, 426
- Power arc, 421
- Power density estimation, 423
- Power dividers, 301
- Power factor, measurement, 334
- Power-line short circuits, 421
- Power measurements (*see* Measurements)
- Power-supply circuits, 206-243
 - d-c, dynamotor, 238
 - transistor, 241
 - vibrator, 239
- Power-supply filter, minimum inductance, 226
- Power-supply transients, 139
- Power-supply voltage control, 213
- Precaution against electric shock, 419
- Predicted-wave Teletype, 314
- Preamplifier, 7, 185, 320, 368
- Pressure drop in air-cooling systems, 272
- Primary power measurement, 333
- Printer margin, 372
- Projected cutoff bias, 145
- Propagation forecasts, 345
- Propagation velocity in cables, 395
- Protective circuit, 260
 - r-f overload, 266
 - speed, 261
- Pulled-crystal keying, 313
- Pulse, peak power measurement, 344
 - sine-squared, 382, 390
 - 2T, 390
- Pulse cross, 387
- Pulse-echo method of fault location, 400
- Pulse-forming network, 198
- Pulse generators, 195
- Pulse measurements, 375
- Pulse modulation (*see* Modulation)
- Pulse techniques, 382
- Pulsed frequency modulation, 200
- Pulser, hard-tube, 197
 - line-type, 197
- Pulsing, grid, 198
- Quadrature circuit, 190
- Quadrature filament operation, 215
- Quadriphase keying, 316
- Quantization noise, 6, 200
- Quantized signal, 193
- Quartz crystals, 12-24
- RA method of TV transmission, 387
- Rad, 425
- Radar, 310
- Radiated power levels, 4
- Radiation, cabinet, 406
- Radiation hazards, 422-426
- Range finder, telegraph printer, 372
- Range transmitter, 309
- Rat race, 298
- Rating factor, 390
- Ratio, carrier-to-noise, 6
 - signal-to-distortion, 177
 - signal-to-noise, 6
- Rayleigh region, 423
- RBE, 425
- RC time-constant chart, 196
- Reactance, variable, cathode-follower, 191
- Reactance charts, 445-447
- Reactance-tube modulator, 190
- Reb, 425
- Receiver, 1, 6
- Rectifier filament, quadrature connection, 215
- Rectifiers, arcback protection, 220
 - current rating, 224
 - gas-filled, 218
 - mercury-vapor, 218
 - semiconductor, 221
 - vacuum, 219
 - xenon, 219
- Recycling, automatic, 251
- Reflection coefficient, 403
- Regulation, power supply, 225
 - r-f driver, 142

- Regulator, a-c voltage, 206
 - d-c voltage, 236
 - electromechanical, 206
 - induction, 208
 - saturable-reactor, 210
 - transformer, resonant, 209
- Rem, 425
- Remote control, 254
- Rep, 425
- Repeaters and translators, broadcast, 328
- Resistance, bleeder, 228
 - nonlinear, 420
- Resolution, radar, 310
- Respiration, artificial, 418
- Response, frequency, of TV amplifier, 375
 - variation with drive level, 381
- Restricted-radiation device, 1
- Resuscitation equipment, 418, 419
- R-f filter, 292
- R-f power density, 422, 423
- R-f radiation hazards, 422-426
- Ring frequency, 325
- Ripple frequency, 216
- RL time-constant chart, 196
- Roentgen, 423
- Roll-off, IRE, 385

- Safe radiation levels, 422
- Safety devices, 419
- Sag, 325
- Satellite transmitters, 318
- Scale, gray, 327
 - in liquid cooling system, 280
- Scale-of-two circuit, 196
- Screen-grid modulation, 148
- Selenium, 426
- Self-balancing bolometer bridge, 341
- Self-balancing calorimetric power meter, 341
- Sensation due to electric current, 417
- Separation of intermodulation noise, 374
- Series-resonant circuit, 20, 111
- Serrasoid modulator, 189
- 75-microsecond preemphasis curve, 368
- Shift, carrier, 142, 362
- Shock, electric, 417-419
- Short circuit, power line, 421
- Short-term stability, 23
- Sideband, disappearance, 183
 - double, 135
- Sideband, monkey chatter, 319
 - significant, 182
 - single (*see* Single sideband)
 - suppression, 170, 178
 - vestigial, 135
- Sine-squared pulse, 382
- Single sideband, 135, 157-178
 - generation, filter method, 158
 - phasing method, 159
 - linearity measurement in amplifiers, 174
 - peak-envelope power measurement, 337
 - signal-to-distortion ratio, 177
 - spectrum conservation, 157
 - speech intelligibility, 318
 - suppression, 178
 - transmitter output rating, 158, 339
 - two-tone test, 175, 363
- Single-stub tuner, 130
- Sliding harmonic method of frequency measurement, 355
- Slotted bridge diplexer, 295
- Slotted line, 396
- Smith chart, 396
- Somatic cell damage, 422
- Space, 3
- Spark gap, 231
- Spark-over distances in air, 450
- Specific gravity and heat of ethylene glycol, 341
- Spectrum analyzer, 134, 195, 358
- Spectrum conservation, 4, 157, 405
- Speech, intelligibility on single sideband, 318
- Spreading effect in crystal filters, 166
- Spurious frequency, 4, 28, 405
- Square waves, differentiation, 195
 - generation, 196
- Stabilitron, 104
- Stairstep signal, 202, 392
- Standard, frequency, primary, 344
 - secondary, 348
- Standard-frequency broadcasts, 347
- Starting sequence, transmitter control, 247
- Step starter, 246
- Stereophonic sound, 320
- Stretch circuit, 327
- Stretcher, line, 301
- Stubs, double, 131
 - single, 130

- Subcarrier, 5, 320
 - color, 325
 - amplitudes, 326
 - phase angles, 326
- Subharmonics, 405
- Supermodulation, 153
- Suppressor-grid modulation, 148
- Surge impedance, 421
- Sweeping oscillators, 378
- Swept-frequency method, fault location, 400
- Swinging choke, 227
- Switch locks, 419
- Switching of transmission lines, 303
- Symmetrical attenuator values, 435
- Sync compression, 385
- Sync pulse, 382
- Synchrocyclotron, 329
- Synchronous code, 314
- Synchronous vibrator power supply, 240
- Synthesizer, frequency, 9, 27
 - AN/ARC-27, 32
 - disciplined oscillator, 27
 - phase-locked variable-frequency oscillator, 33
- T network, 120, 448
- T-pad values, 435
- Tacan, 310
- Teflon vapor, 426
- Telegraph distortion, 372
- Telemetering, transmitter remote control, 257
- Telemetry, 190, 318
- Teletype, distortion measurements, 373
 - phase-shift, 314
 - predicted-wave, 314
 - single-sideband techniques, 314
- Television, amplifiers, linearity, 391
 - clamp failure, 385
 - differential phase, 392
 - "funny-paper" effect, 389
 - measurements (*see* Measurements)
 - picture monitor, 387
 - sideband analyzer, 376
 - sync compression, 385
 - transmitter, 320
 - waveform display, 382
- Television signal, bandwidth, 321
 - black negative, 383
 - Television signal, color, 326
 - monochrome, 375
 - multiburst test, 378
 - polarity, 383
 - vertical-interval test, 392
- Temperature compensation, 11
- Temperature measurement of tube envelopes, 281
- Termination wattmeter, 340
- Ternary code, 194
- Tetrodes, modulation of, 143
 - neutralization of, 66
 - screen-grid supply, 86
 - self-neutralizing frequency, 69
- Thermal circuit breaker, 263
- Thermistor, 341
- Thermocouple ammeters, 336
- Thévenin generator, 379
- 30-day moving average, WWV frequency, 345
- Three-wattmeter measurement of three-phase power, 334
- Threshold extension, 6
- Thyrite protection, 223, 420
- Tilt, 325
- Time-constant chart, 196
- Time-division multiplex, 315
- Time domain, 182
- Time measurements, 360
- Time signals, 345
- Timing capacitor, 240
- Toll circuit noise, 374
- TR switch, 312
- Transfer oscillator, 350, 366
- Transformers, 228
 - buck-boost, 214
 - insulating oils, 229, 426
 - insulation, 228
 - utilization, 229
- Transient, keying, 138, 313
 - measurement techniques, 382
 - in power-supply filter, 139
 - protection, 222, 229
- Transient response of TV system, 324
- Translator, 328
- Transmission, vestigial-sideband, 153
- Transmission level, zero, 374
- Transmission line, attenuation, 342
 - cutoff frequency, 406
 - fault location in, 400
 - lumped circuit equivalents, 130

- Transmitters, aircraft, 318
 - amateur, 316
 - avigation, 309
 - broadcast, 319
 - emission characteristics, 2
 - f-m, 319
 - glide-path, 309
 - hazards associated with, 417
 - localizer, 309
 - marine, 317
 - mobile, 317
 - navigation, 309
 - pulse, 316, 318
 - radar, 310
 - satellite, 318
 - Tacan, 310
 - telegraph, 312
 - telemetry, 318
 - telephone, 316
 - television, 320
 - translator, 328
 - unlicensed, 1
- Transmitting tube types, 95
- Trapezoidal pattern, 363
- Treatment for electric shock, 418
- Triple-stub tuner, 131
- TS-175/U frequency meter, 356
- Tube, power, 89-107
 - traveling-wave, 12
- Tuned amplifier, 380
- Tuning, automatic, 259
- TVI, 406
- 2T pulse, 390
- Two-tone test signal, 177, 363
- Two-wattmeter method of three-phase power measurement, 334
- Universal curves for L, pi, and T networks, 448, 449
- Unlicensed transmitters, 1
- UT-2 time, 345
- Variable-frequency oscillator, crystal-controlled, 22
 - inductance-capacitance, 10
- Varistor, 420
- Vectorscope, 327
- Velocity of propagation in transmission lines, 395
- Velocity modulation, 98
- Vernier method of time measurement, 360
- Vestigial-sideband filter, 293
- Vestigial-sideband transmission, 153
- Vibrator power supply, 239
- Video modulator, 147
- Video transmission set, 377
- Voltage control, power supply, 213
- Voltage-standing-wave ratio (VSWR), 394, 396, 403
- Voltage stress in tubes, 266
- Volume unit, 335
- Vu meter, 335
- Wattmeter, absorption, 340
 - termination, 340
- Waveforms, television, 382
- Waveguide, arc detection in, 267
 - cutoff frequency, 401
 - fault location in, 400
 - impedance matching in, 132
- Wavemeters, 357
- Weatherstripping, electronic, 406
- White compression, 385
- White stretch, 327
- Wire, current-carrying capacity, 439
- Wire tables, 438
- Wireless microphone, 2
- Words per minute to bauds or cps, conversion factors, 3
- WWV schedules, 345-347
- WWVH schedules, 345-347
- WWVL schedules, 347
- Wye connection, 335
- X-ray exposure, permissible, 425
- X-rays, 422
 - radiation hazards, 422-426
- Zero transmission level, 374